

Multiconductor Transmission Line Modeling of Crosstalk Between Cables in the Presence of Composite Ground Planes

Jesper Lansink Rotgerink , Ramiro Serra, and Frank Leferink , *Fellow, IEEE*

Abstract—Modern transportation systems, such as aircraft, are increasingly replacing metal body parts for composite materials, such as carbon-fiber reinforced plastics. Despite the multiple advantages in terms of weight, size, and fuel consumption, this trend is posing a challenge for protection of cables against electromagnetic interference. Early risk assessment and optimization of cable design in modern aircraft require reliable methods that can quickly and accurately estimate crosstalk behavior in the presence of lossy ground planes. This article presents two novel methods to incorporate such lossy ground planes into the crosstalk modeling of cable bundles. The first method considers the ground plane as a discrete collection of cylindrical conductors. In the second method a ground impedance matrix is derived for lossy ground planes with a finite thickness. Results of both methods are compared to full-wave simulations and measurements, yielding excellent results in terms of accuracy and computation times. The discretized ground plane method is also applied to the case of wire pairs that are enclosed by two ground planes, both aluminum and carbon-fiber reinforced plastic, as a first step towards investigation of wiring that is embedded in thermoplastic material. Once more simulations and measurements are in good agreement.

Index Terms—Crosstalk, carbon-fiber reinforced plastics (CFRP), composites, multiconductor transmission lines, lossy ground planes.

I. INTRODUCTION

SUSTAINABLE flight is one of the most prominent objectives of current developments in aviation industry. To reach that goal there is much focus on the development of more electric aircraft (MEA), including hybrid or full electrical propulsion. As a consequence, there is a significant increase in

the amount of on-board electrical wiring. This highlights the desire for an electrical wiring interconnection system (EWIS) on-board aircraft that is optimized in terms of weight, volume, and costs, while keeping compliance with safety and electromagnetic compatibility regulations. As MEA will implement a significantly high number of power electronic converters and sensors, crosstalk between cables is an essential topic.

Transmission lines (TL) are broadly covered in literature, especially when TLs are placed in free space or above conducting ground planes. The multiconductor TL (MTL) model developed by Paul [1] is widely accepted for crosstalk analyses. Paul also includes losses in the per-unit-length (p.u.l.) resistance matrix of his MTL formulations. However, he clearly states the shortcomings regarding the distribution of return currents in large, lossy ground planes. It is well-known that it is good practice to place cables close to a good conducting surface, as this generally reduces crosstalk levels [2], [3]. However, in modern aircraft more and more metal parts are replaced by composite materials such as carbon-fiber reinforced plastics (CFRP) [4], [5]. As a consequence, there is a need to model the effects on crosstalk behavior due to replacement of metal parts in aircraft by less conducting material such as CFRP. Thus, simulation models for crosstalk in cable bundles that can include the lossy properties of CFRP ground planes are a requirement for modern aircraft EWIS design.

Already in 1926, Carson [6] investigated the influence of lossy soil on the behavior of the transmission lines in a paper on wave propagation in overhead lines. More specifically, he derived formulations, which describe the distribution of electrical currents in a lossy ground, as well as an expression for the mutual impedance between two transmission lines that have a lossy ground as return. Since then, much research has focused on simulation models for transmission lines above lossy earth. Improvements to Carson's formulation have been made [7]–[10], and papers have been published on the use of these formulations in transient analysis of overhead lines above lossy grounds [11]–[13]. Rachidi provides an extensive overview of field-to-TL coupling models that can also include the effects of lossy earth [14]. However, most papers focus on a lossy ground with infinite thickness (e.g., soil), as well as the transient (time-domain) analysis of transmission lines under the influence of incident electromagnetic (EM) fields and lightning strikes [14], [15].

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The focus in this article is on frequency domain simulations of crosstalk between cables in the presence of ground planes with finite conductivity and finite thickness. In [16], it was observed that skin depth and ground plane thickness are the main factors dictating a shift in crosstalk behavior in the frequency domain. In the higher frequency range crosstalk in the presence of a CFRP ground plane is similar to that above good conducting ground, while for lower frequencies crosstalk levels are hardly affected by the presence of CFRP. These observations in [16] were obtained from full-wave simulations and confirmed by measurements. Most commercially available full-wave solvers can model the effects of CFRP on crosstalk. However, such methods are computationally expensive and faster methods are desired, especially for industrial applications. The MTL equations are a widely used and much quicker method. An accurate and broadband numerical approach to estimate the p.u.l. parameters that includes skin effect, current crowding and semiconducting material is presented in [17]. However, this still requires the numerical method of moments (MoM) to solve the p.u.l. parameters.

Time-efficient broadband simulation methods that can provide an accurate first estimate of crosstalk in the presence of lossy ground planes are very useful for early risk assessment in EWIS design. Moreover, such methods can be used in cable bundle optimization, as well as in sensitivity analysis with respect to designable parameters. This article proposes two novel methods to incorporate lossy ground planes, including corresponding skin and proximity effects, directly into the MTL equations. Both methods are efficient solutions that are easy to apprehend and implement. The first method makes use of a discretization of the ground plane into cylindrical conductors. In [18], such a discretization of a copper ground plane was used to analyze ground currents. The second method of this article utilizes a ground impedance matrix. This impedance matrix is derived from an expression for mutual impedance of overhead lines above stratified earth [9]. Both methods are applied to crosstalk between wire pairs above a CFRP or aluminum ground plane. Moreover, the method of discretized ground planes is also applied to cabling between two CFRP or aluminum ground planes. Results of these simulations are compared to full-wave MoM simulations with Feko [19] and to measured crosstalk.

The rest of this article is organized as follows. Section II of this article discusses the two transmission line models that include lossy ground planes. In Section III, simulation results of two cases are shown and compared to full-wave simulations and measurements. Finally, Section IV concludes this article.

II. TRANSMISSION LINE MODELS

Broadly used simulation methods to analyze crosstalk between cables are based on the MTL

$$\frac{d}{dz}\mathbf{V}(z) = -\mathbf{Z}\mathbf{I}(z), \quad \frac{d}{dz}\mathbf{I}(z) = -\mathbf{Y}\mathbf{V}(z). \quad (1)$$

In these equations \mathbf{V} and \mathbf{I} are n -dimensional vectors representing the voltages and currents in each of the n nonreference conductors of the MTL along its propagation direction z . For frequency domain solutions, the p.u.l. impedance matrix \mathbf{Z} and

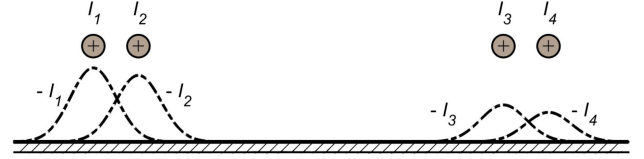


Fig. 1. Illustration of (return) current distribution in a lossy ground plane.

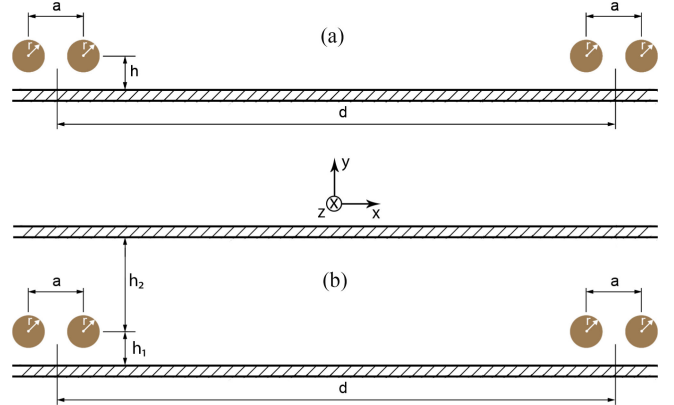


Fig. 2. Illustration of cross section of test case with two wire pairs (a) above a single ground plane and (b) between two ground planes.

admittance matrix \mathbf{Y} are usually defined by

$$\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L} \quad \mathbf{Y} = \mathbf{G} + j\omega\mathbf{C}. \quad (2)$$

Here, ω is the angular frequency. The resistance matrix \mathbf{R} , the capacitance matrix \mathbf{C} , the inductance matrix \mathbf{L} , and the conductance matrix \mathbf{G} , together form the p.u.l. parameters, which contain all cross-sectional information of the MTL.

Equations (1) and (2) describe the evolution of the voltages and currents along the line. All voltages are defined with respect to the reference conductor and the return current of all conductors also flows in this reference conductor. For a perfectly electric conducting (PEC) ground plane this return current is concentrated in a delta-peak directly below the conductor itself. However, if losses are present in the ground, these currents tend to spread out due to dispersion [1], [6], as is shown by the illustration in Fig. 1. If we extend this further towards the case of no ground plane, the return currents will have vanished. Therefore, to model a CFRP ground plane with transmission line equations, we need to accommodate losses in the reference conductor. This article presents two models, which can account for (return) current distributions in lossy ground planes. One method discretizes the ground plane by a series of lossy wires, and the second implements a ground impedance matrix. In the following sections, the two methods are described and explained based on the test case in Fig. 2(a). Two wire pairs are separated by a distance $d = 20$ mm and are at a height h above a ground plane with thickness t_g . This ground plane is considered to be made of either aluminum with $t_g = 1.5$ mm or CFRP with $t_g = 1.3$ mm. The intrapair separation a and the wire radius r will be equal to 2.2 mm and 0.55 mm, respectively, throughout this article. Both wire pairs are assumed to be terminated with a differential-mode (DM) resistance $R_d = 100 \Omega$ at both sides. Fig. 2(b) also shows a case in which a second ground plane is

placed above the wire pairs. This test case will be analyzed in Section III.

Assume a DM source is included in the terminations of the culprit wire pair. Consequently, a voltage will be induced in the terminations of the victim wire pair. Then, differential-mode near-end crosstalk (NEXT) $\gamma_{NE,DM}$ can be defined as

$$\gamma_{NE,DM} = \frac{\mathbf{U}_2^T \mathbf{V}_0}{\mathbf{U}_1^T \mathbf{V}_0}. \quad (3)$$

Here, \mathbf{V}_0 is the vector of which entry i represents the voltage of conductor i at the source side of the MTL with respect to the reference. \mathbf{U}_1 and \mathbf{U}_2 are vectors that are used to obtain the correct combination of conductor voltages in the numerator and denominator. For two wire pairs above a single infinite ground plane these would be defined as

$$\mathbf{U}_1 = (-1, 1, 0, 0)^T, \quad \mathbf{U}_2 = (0, 0, -1, 1)^T. \quad (4)$$

Far-end crosstalk (FEXT) can be defined in a similar way by using the voltages and currents at the other side of the transmission line. This article focusses only on NEXT, since the methods and models are analogously applied to FEXT.

Equation (3) shows that by solving the vector of voltages at the source side of the TL, crosstalk can be computed. These voltages can be solved by (2)–(6) in [2], depending on the termination network representation. In that solution, the p.u.l. matrices \mathbf{C} , \mathbf{L} , and \mathbf{R} are required. The conductance \mathbf{G} is assumed to be negligible. For an infinite PEC ground plane the approximate logarithmic expressions for p.u.l. capacitance and inductance as given in [1] can be used. Moreover, in the two new methods presented in the following two sections that include lossy ground planes, for the self and mutual capacitances and inductances these same expressions can be used. The major differences that will be introduced are in resistance matrices and/or the termination impedance matrices.

A. Discretized Ground Plane

The first solution presented in this article is an MTL model with discretized ground plane (MTL-DG). In this method, the ground plane is modeled as a series of cylindrical conductors parallel to the z -axis. The conventional MTL equations given by (1)–(2) can be used, however, the dimensions of all matrices are increased with the number of conductors in the ground plane n_g . Moreover, the resistance matrix and termination matrices have to be adapted. Thus, to model the CFRP ground plane in the configuration of Fig. 2(a), instead of an infinite ground plane a finite ground plane is introduced, which will be represented by an array of n_g adjacent cylindrical conductors. An example with a single layer of ground conductors, each with diameter t_g , is shown in Fig. 3. In general, multiple layers of ground conductors can be used, as long as the diameters add up to t_g . Since in the measurements the ground planes are very large compared to the separation between the wire pairs, an infinite ground plane can be assumed. Therefore, in this article, n_g is also chosen large enough to mimic an infinite ground plane. Simulations to confirm this follow in Section III. To avoid confusion, in the following the conductors of the ground plane will be referred

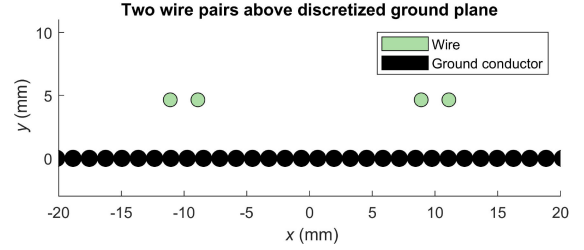


Fig. 3. Illustration of (zoomed-in) cross-section for a cable configuration with two wire pairs above a discretized ground plane.

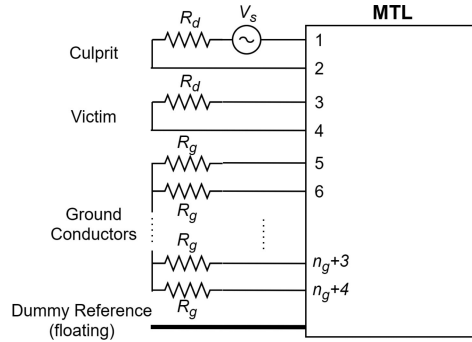


Fig. 4. Termination schematics at the source side for the two wire pairs, terminated with DM resistance R_d , above a discretized ground with n_g ground conductors that are connected to each other by the resistances R_g . Culprit terminations include a voltage source V_s .

to as ground conductors, while the four conductors in the wire pairs will be named wires.

With the replacement of the infinite ground plane by the (finite) array of ground conductors, there is no more natural choice for the reference conductor since, in MTL theory, the reference conductor carries a net current equal to the sum of the return currents from all other conductors. To resolve this issue, a “dummy” reference conductor is introduced far away from the wires and ground plane under consideration, in this article at $x = 0$ m and at $y = -100$ m. Moreover, the termination network (shown in Fig. 4) is designed to have all conductors floating with respect to the dummy reference. In calculations this dummy reference appears just as a placeholder for the voltage integrals, but it is a dead conductor. Due to the distance to the structure and termination impedances that approach infinity due to the floating dummy reference, the dummy carries no net (return) current and has no effect to the local behavior of the wire pairs above the ground. All currents flow locally in the wire pairs and the ground plane. Since the cross section of the actual cabling remains small compared to the wavelength, TL theory can be applied [20]. The two wire pairs are terminated on both sides by the DM resistance $R_d = 100 \Omega$. To have all wires and ground conductors floating with respect to the dummy reference conductor, a Norton equivalent representation is required to incorporate the termination network into the current implementation of the MTL model. In theory, the ground conductors are short circuited. However, this would create a singularity in the admittance matrices introduced in (5) below. To avoid this limitation, a small resistance $R_g = 1 \text{ m}\Omega$ (shown in Fig. 4) is used to mimic a short between all ground conductors. The value of R_g is chosen such that it

is negligible with respect to the losses in a ground conductor, which for a single layer of ground conductors equals 36Ω at 0.1 MHz [computed by (9)]. In this way, R_g does not affect crosstalk results and allows for current distribution through the ground plane. This results in the following admittance matrices at source and load side \mathbf{Y}_S and \mathbf{Y}_L :

$$\begin{aligned} \mathbf{Y}_{S/L} &= \begin{bmatrix} \mathbf{Y}_N & \mathbf{0} \\ \mathbf{0} & \mathbf{Y}_{\hat{g}} \end{bmatrix}, \quad \mathbf{Y}_N = \begin{bmatrix} \mathbf{Y}_c & \mathbf{0} \\ \mathbf{0} & \mathbf{Y}_v \end{bmatrix}, \\ \mathbf{Y}_k &= \begin{bmatrix} R_{d,k}^{-1} & -R_{d,k}^{-1} \\ -R_{d,k}^{-1} & R_{d,k}^{-1} \end{bmatrix} \\ \mathbf{Y}_{\hat{g}} &= \frac{1}{\left(1 + \frac{1}{n_g - 1}\right) R_g} \mathbf{1}_{n_g} - \frac{1}{n_g R_g} \mathbf{1}'_{n_g}. \end{aligned} \quad (5)$$

Here, \mathbf{Y}_N contains the wire pair terminations and $\mathbf{Y}_{\hat{g}}$ the ground conductor terminations, $\mathbf{1}_m$ represents the $m \times m$ identity matrix, and $\mathbf{1}'_m$ the matrix with ones on all nondiagonal elements. The subscript k can be c for culprit and v for victim.

To include the ground current distribution and skin effect in the MTL model the analytical resistance of cylindrical conductors with finite conductivity is included in the resistance matrix for all ground conductors. This yields a p.u.l. impedance matrix equal to

$$\mathbf{Z} = \mathbf{R}_w + \mathbf{Z}_g + j\omega \mathbf{L}_w \quad (6)$$

in which matrix \mathbf{L}_w contains the self and mutual inductances of the entire MTL (including far-away reference). The elements of this inductance matrix, as well as the elements of the capacitance matrix, are computed by the following analytical expressions [1]:

$$l_{ii} = \frac{\mu_0}{2\pi} \ln \left(\frac{d_{i0}^2}{r_i r_0} \right) \quad l_{ij} = \frac{\mu_0}{2\pi} \ln \left(\frac{d_{i0} d_{j0}}{d_{ij} r_0} \right) \quad \mathbf{C} = \mu_0 \varepsilon_0 \varepsilon_r \mathbf{L}^{-1}. \quad (7)$$

Here, ε_0 and μ_0 are the free space permittivity and permeability, ε_r is the relative permittivity of the surrounding medium, and d_{ij} is the distance between conductors i and j , where subscript 0 refers to the dummy reference. The conductor radius r_i is either the wire radius (r in Fig. 2) or the ground conductor radius. The radius of the dummy reference r_0 is chosen equal to the wire radius. In (6), the matrix \mathbf{R}_w contains the resistances of the wires that form the two wire pairs. However, in this article, the resistances in the wires are neglected, resulting in $\mathbf{R}_w = \mathbf{0}$. Finally, \mathbf{Z}_g contains the resistances and inductances of the conductors that form the ground plane, for which the formulations described by Paul are used [1]

$$\mathbf{Z}_g = z_g \mathbf{1}_{n_g}, \quad z_g = r_g + j\omega l_g. \quad (8)$$

Here,

$$\begin{aligned} r_g &= \frac{2q}{\sigma_g d \pi t_g^2} \left(\frac{\text{ber}(q) \text{bei}'(q) - \text{bei}(q) \text{ber}'(q)}{(\text{bei}'(q))^2 + (\text{ber}'(q))^2} \right) \\ l_g &= \frac{4\mu_0}{8\pi q} \left(\frac{\text{bei}(q) \text{bei}'(q) + \text{ber}(q) \text{ber}'(q)}{(\text{bei}'(q))^2 + (\text{ber}'(q))^2} \right), \quad q = \sqrt{2} \frac{d_g}{2\delta}. \end{aligned} \quad (9)$$

Here, ber and bei are Kelvin functions, d_g is the ground conductor diameter, which for the discretization in Fig. 3 is equal to t_g , and δ is the skin depth, given by

$$\delta = 1/\sqrt{\pi f \mu_0 \sigma_{gd}}, \quad \sigma_{gd} = 4\sigma_g/\pi. \quad (10)$$

Here, σ_g is the conductivity of the ground plane. The value σ_{gd} is the conductivity used for the ground conductors, which contains a correction factor to account for the area that is missing since we are discretizing a rectangular area by circles.

B. Ground Impedance

The second method to incorporate the losses and current distribution in the transmission line models is to include a ground impedance matrix into the MTL equations (MTL-GI). The TL model remains as given by (1) and (2). In the MTL-GI method, as opposed to the MTL-DG method, the ground plane can be taken as infinite. Therefore, the configuration of Fig. 2(a) results in a 4×4 inductance and capacitance matrix of which the elements are computed by [1]

$$l_{ii} = \frac{\mu_0}{2\pi} \ln \left(\frac{2h}{r} \right) \quad l_{ij} = \frac{\mu_0}{4\pi} \ln \left(1 + \frac{4h^2}{d_{ij}^2} \right) \quad \mathbf{C} = \mu_0 \varepsilon_0 \varepsilon_r \mathbf{L}^{-1}. \quad (11)$$

The conductance matrix is still assumed to be zero and the p.u.l. impedance matrix is again defined as given in (6), this time resulting in a 4×4 matrix. Analogous to the MTL-DG method, we neglect the resistances in the wires. The effects of CFRP should, now, be included in \mathbf{Z}_g . There is extensive literature about such impedance matrices that include the losses of soil into transmission line equations. Carson [6] was one of the first ones to derive an equation for ground impedance in which the soil was represented by a lossy half-space. By adapting the formulation for ground impedance of a stratified soil that is given in [9] to the case of a ground plane with finite thickness and conductivity, the ground impedance matrix for MTL-GI is given by

$$[\mathbf{Z}_g]_{ij} = \frac{j\omega\mu_0}{\pi} \int_0^\infty \frac{\cos(\alpha(x_i - x_j)) e^{-\alpha(h_i + h_j)}}{\alpha + A} d\alpha. \quad (12)$$

Here,

$$\begin{aligned} A &= n_2 \left[\frac{n_1 - n_2 + (n_1 + n_2) e^{2t_g n_2}}{n_2 - n_1 + (n_1 + n_2) e^{2t_g n_2}} \right] \\ n_1 &= \sqrt{\alpha^2 + j\omega\mu_0}, \quad n_2 = \sqrt{\alpha^2 + j\omega\mu_0\sigma_g}. \end{aligned} \quad (13)$$

Here, x_i gives the x position of wire i , while h_i is the height of wire i above the ground plane. Apart from this adaptation, the MTL equations remain unchanged and the termination admittance matrices are defined by

$$\mathbf{Y}_S = \mathbf{Y}_L = \begin{bmatrix} \mathbf{Y}_c & \mathbf{0} \\ \mathbf{0} & \mathbf{Y}_v \end{bmatrix}. \quad (14)$$

III. RESULTS

In this section, crosstalk results from the two MTL models presented in Section II are compared to full-wave simulations obtained with Feko and to measurements. The comparison is performed for two configurations. At first, the configuration of

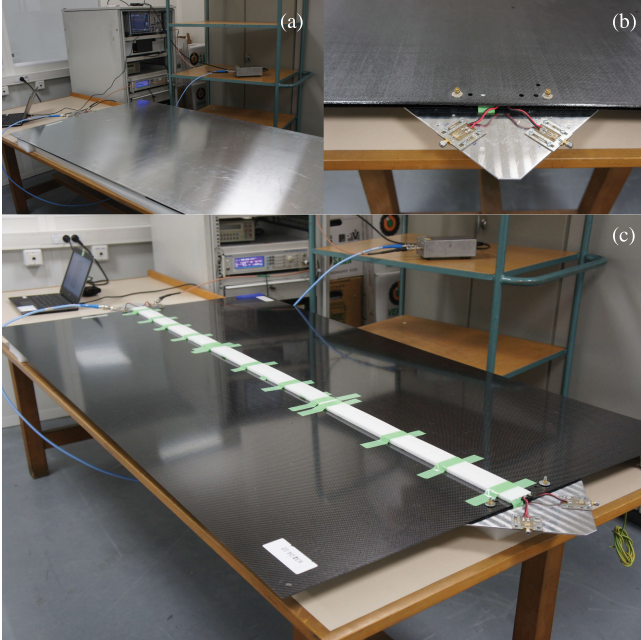


Fig. 5. Pictures of measurement setup: crosstalk (a) between aluminum ground planes - overview of measurement setup with spectrum analyzer, generator and preamplifier, (b) between two CFRP ground planes - close-up of termination to baluns, and (c) above a single CFRP ground plane.

two wire pairs above a single ground plane, as shown in Fig. 2(a), is computed by both the MTL-DG and the MTL-GI method and compared to Feko simulations and measurements. Thereafter, the MTL-DG method is applied to a configuration with wire pairs between two ground planes, shown in Fig. 2(b). The MTL-DG method can be readily adapted to this second case, due to its flexibility and ease of use. It only requires the inclusion of a higher number of ground conductors. The MTL-GI method is not directly applicable to this case with two ground planes, unless for instance a Laplace solver is incorporated and (12) is modified accordingly.

All measurements are performed with the combination of a signal (tracking) generator as source and a spectrum analyzer to measure the victim voltage (see Fig. 5). The balanced wire pairs are connected to the unbalanced measurement equipment using baluns. These baluns can be used in the frequency range of 100 kHz–440 MHz. Their frequency-dependent behavior is calibrated out of the measurements. Rohacell spacers with low loss and relative permittivity nearly equal to 1 are used to ensure fixed separation distances and heights of the ground plane along the length of the transmission line, which equals 1.8 m. The ground planes used in the measurements were 1.8 m in length and 1 m in width. Their thickness equaled 1.3 mm and 1.5 mm for the CFRP and aluminum grounds, respectively. At the end of the transmission lines, the wire pairs are separated from each other and connected to the balun boards [see Fig. 5(b)]. Connections between the baluns and the ground planes are established by screws through the ground plane, completed with conductive glue for tighter fastening of the screws. This results in a common-mode impedance of 450 Ω , which is high enough

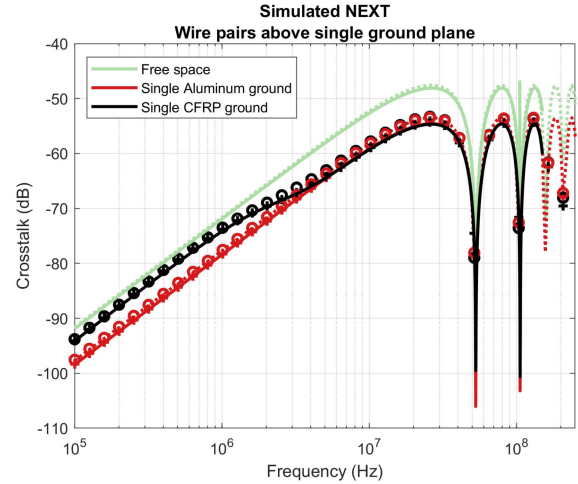


Fig. 6. Simulated NEXT between two wire pairs in three different configurations. The solid lines are obtained with Feko, while the dotted lines are with the analytic MTL equations. The circles give results computed by MTL-GI, while the plus signs show results for the MTL-DG method.

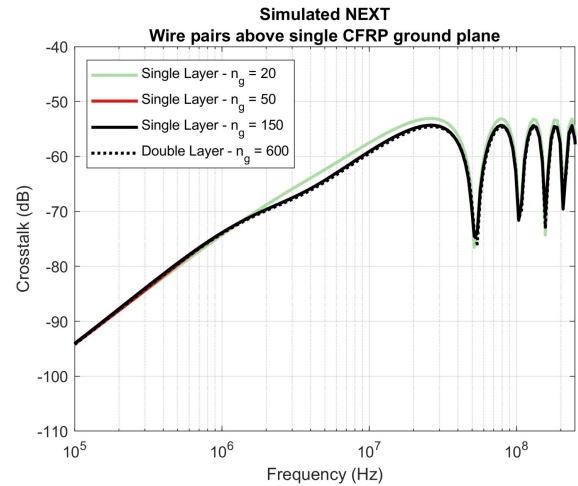


Fig. 7. Simulated NEXT (MTL-DG) for wire pairs above a single CFRP ground plane. Comparisons are made between three different sizes of a single layer ground plane and a double layered ground plane.

to use in simulations only DM terminations [16]. In simulations, the CFRP is modeled with an isotropic equivalent conductivity, which was found to be valid in [16], [21], and [22]. For our CFRP planes, this equivalent conductivity equals 16.500 S/m [16].

In simulations, a homogeneous dielectric with $\epsilon_r = 2.5$ is assumed to account for wire insulation and foam spacers. Feko simulations are performed by making use of an isotropic thin dielectric layer to which CFRP properties are asserted.

A. Single Ground Plane

Figs. 6 to 9 show results for crosstalk between wire pairs in free space, as well as for wire pairs at a height $h = 4$ mm above a ground plane [cross section of the latter shown in Fig. 2(a)]. Fig. 6 shows a comparison between near-end crosstalk computed by various simulation models. Results are

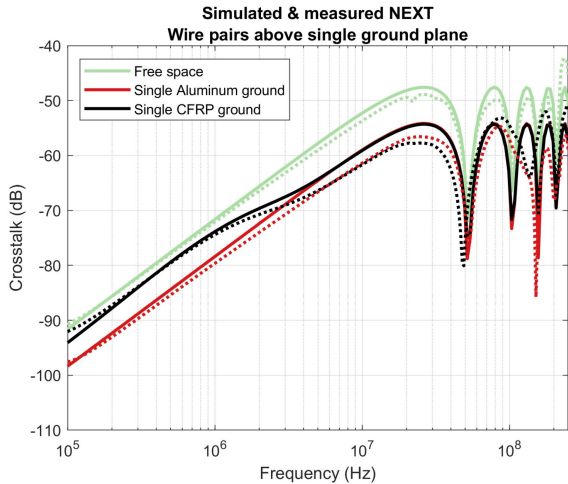


Fig. 8. Comparison between measured and simulated NEXT for wire pairs in free space, or above a single ground plane. Solid lines are results of MTL simulations (also shown in Fig. 6, analytic MTL for free space and MTL-GI for the ground plane cases). The dotted lines are measured crosstalk results.

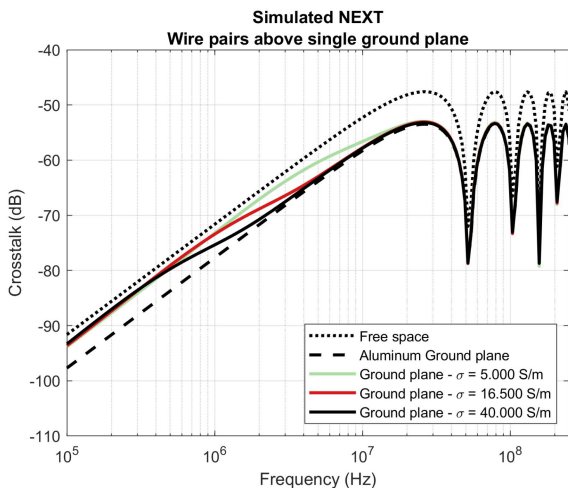


Fig. 9. Simulated (MTL-DG method) NEXT between two wire pairs above a single ground plane with different conductivities. MTL results for free space and PEC ground plane are given as reference.

given for three configurations: wire pairs in free space, wire pairs above an aluminum ground plane and wire pairs above a CFRP ground plane. All three configurations are simulated with Feko and the results are shown by the solid lines. With classic MTL simulations, in which the logarithmic (wide-separation) expressions by Paul are used for inductance and capacitance, the dotted lines are obtained.

The results in Fig. 6 confirm the well-known phenomenon that crosstalk decreases for cables that are placed close to a metallic ground plane, when compared to wire pairs in free space. For these two cases MTL results are equal to those generated by Feko. The MTL-DG method introduced in the previous section yields levels shown by the plus signs in Fig. 6. With this method both the aluminum ground plane case and the CFRP ground plane case are shown and results show a perfect match to those

obtained with Feko. Finally, results obtained with MTL-GI are given by the circles. Again, these results match perfectly to the other simulation methods.

Various discretization levels might be used in the MTL-DG method. Fig. 7 shows a comparison of results for a CFRP ground plane with a single layer and three different values for n_g , as well as a discretization with two layers of wires. In all cases, the ground conductors are touching. Increasing n_g in a single layer will therefore increase the ground plane size. Introducing a second layer of ground conductors might increase accuracy, but also increases computation time. From Fig. 7, it can be concluded that a single layer with 50 ground conductors is sufficient to represent an infinite ground plane, for the case under consideration. Therefore, in all MTL-DG simulations in this article a single layer with $n_g = 50$ is used.

Concluding, similar results are obtained by all three simulations methods. However, simulation times are very different. Fine meshing is needed on a ground plane below a transmission line in order to accurately represent the characteristic impedance in Feko. Therefore, simulations with Feko depend on the wire height and vary for $h = 4$ mm to 1 mm from 60 s to 1060 s for a single-frequency computation, when using 10 cores on a Linux server (Intel(R) Xeon(R) Gold 6148 CPU @ 2.40 GHz, 80 cores available). At the same time, the MTL-DG and MTL-GI methods take only 6.5 and 3 s, respectively, for computing 200 frequencies on a simple laptop (1 core on Intel i5 processor, 8GB RAM). Therefore, simulation time for a single frequency computation is improved by a factor of 1900–70 000. When comparing the two novel MTL methods, MTL-GI is slightly more time-efficient and enables actual infinite ground plane simulation. However, as illustrated also by the case in Section III-B, the MTL-DG method provides more flexibility.

Fig. 8 shows the comparison of the MTL simulations with measured NEXT. For the ground plane cases, the MTL-GI method is shown. Some uncertainties in the measurement setup cause the measured crosstalk curve have a slope slightly less than the expected 20 dB/dec., which causes slight differences in the 0.1–0.3 MHz range. However, in general the comparison between simulations and measurements is good. The simulation results clearly show that CFRP affects crosstalk behavior as stated in [16]. There is a clear distinction between low and high-frequency behavior, in which both the conductivity and thickness of the ground plane are important parameters. The distinction is determined by the field penetration in the CFRP material [23]. At low frequencies, for which the skin depth is much larger than the thickness of the ground plane, the electromagnetic fields tend to distribute throughout the ground plane and consequently the currents distribute in depth and transversal directions of the ground. At high frequencies, for which the skin depth is small compared to the ground plane thickness, the fields penetrate only the top part of the ground plane given by one skin depth and currents concentrate at the surface and below the wires. For the good conducting aluminum ground planes this high-frequency effect holds throughout the entire frequency range considered in this study.

These observations are confirmed by results shown in Figs. 9 and 10. Fig. 9 shows that for lower conductivity, the transition

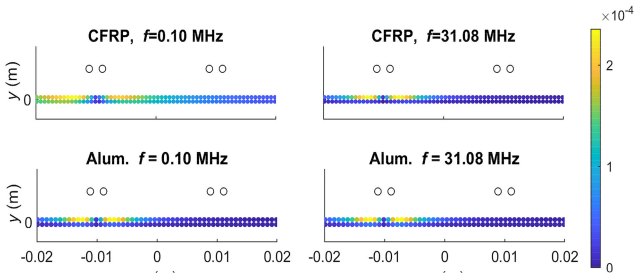


Fig. 10. Simulated currents (in ampere) at the near-end side of all conductors in the ground plane at 2 different frequencies, for both CFRP and aluminum.

from free-space-like to PEC-like behavior occurs at a higher frequency than for higher conductivity. For the conductivity of our measured CFRP, i.e., 16.500 S/m, Fig. 10 also shows current distributions in the ground plane. To better show current distributions, a ground plane discretization with two layers of ground conductors is used for the computation of results in Fig. 10. A big advantage of the MTL-DG method is that these current distributions are readily available after solving the MTL equations. For all subplots the left wire pair is the culprit, and the magnitudes of ground plane currents are shown in correspondence with the color bar. The three upper subfigures of Fig. 10 show the current distributions in a CFRP ground plane, while the three bottom subfigures are for an aluminum ground plane. Indeed, good field penetration in the CFRP for the lower frequencies causes the currents to distribute wider across the ground plane, while they become more concentrated below the current-driving wires for higher frequencies, showcasing proximity effects. In aluminum, the currents in the ground plane are concentrated below the wires in the entire frequency range considered in this study.

B. Double Ground Plane

To further illustrate the capabilities of the proposed MTL-DG method, this section shows results of crosstalk between wire pairs between two ground planes. The cross section is illustrated in Fig. 2(b). Such a configuration can be interesting as a step towards simulation of integrated or embedded wiring, which is of interest to the aviation industry.

Fig. 11 shows simulated NEXT results for two wire pairs between two ground planes, with $h_1 = h_2 = 4$ mm. As reference, the free space results are also shown. Clearly, crosstalk is even further reduced when a second ground plane is placed above the wire pairs. For a single ground plane, the decrease was roughly 8 dB compared to the free space case. From Fig. 11, we see that introduction of a second ground plane decreases crosstalk with another 15 dB, for the case under study. Moreover, it can be observed that the general behavior of crosstalk between CFRP ground planes shows the same tendency as crosstalk above a single CFRP ground plane, but with a different level. Again, for low frequencies the crosstalk is similar to that of wire pairs in free space. Fig. 11 shows that the agreement between measurements and MTL simulations are very good for all three cases. The results of the MTL-DG method and Feko show a perfect match

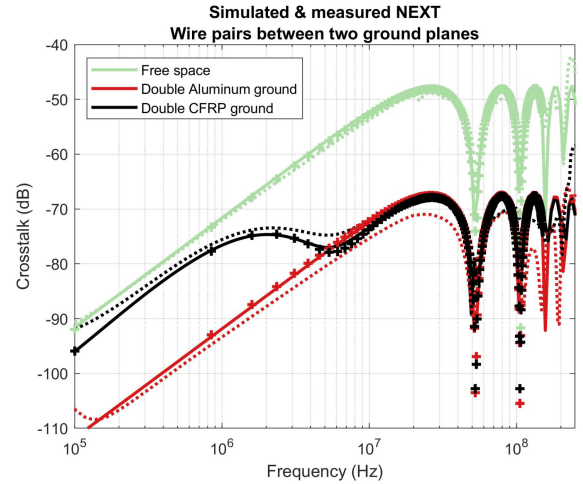


Fig. 11. Comparison between simulated (MTL-DG – solid lines, Feko – markers) and measured NEXT (dashed lines) between two wire pairs between two ground planes.

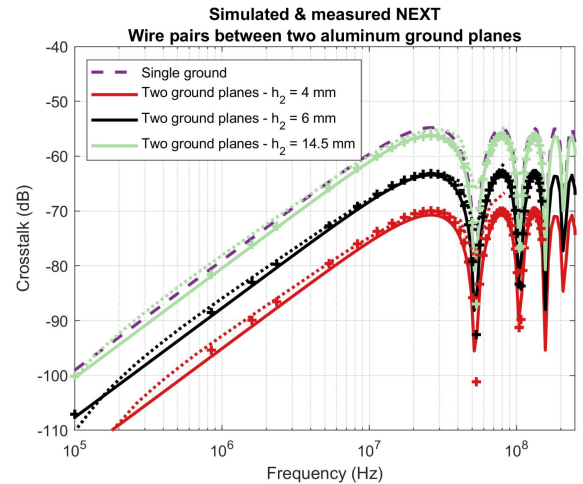


Fig. 12. MTL-DG (solid lines), Feko (markers) and measured (dotted lines) NEXT for two wire pairs between two aluminum ground planes. The lower ground plane is at 3.5 mm, while the height of the other ground plane is changed. MTL-DG result for single aluminum ground plane is shown as reference (purple dashed line).

for all three cases. However, for the case of two ground planes, the MTL-DG method requires only 40 s to run 200 frequencies on a laptop, while Feko computation times vary for $h = 4$ mm to $h = 1.1$ mm from 180 to 14.100 s, to run a single frequency on 10 cores of our Linux server. This implies a factor 900–70 000 simulation time improvement of the MTL-DG method, when compared to Feko.

Finally, Figs. 12 and 13 show results of MTL-DG and Feko simulations, and measurements for two wire pairs between two aluminum and two CFRP ground planes. The lower ground plane is kept 3.5 mm from the wire pairs, while the distance of the upper ground plane is varied. For all distances, MTL-DG and Feko again show a perfect match. Also the simulations and measurements coincide very well. Only for the CFRP ground planes slight differences occur, which are most likely caused

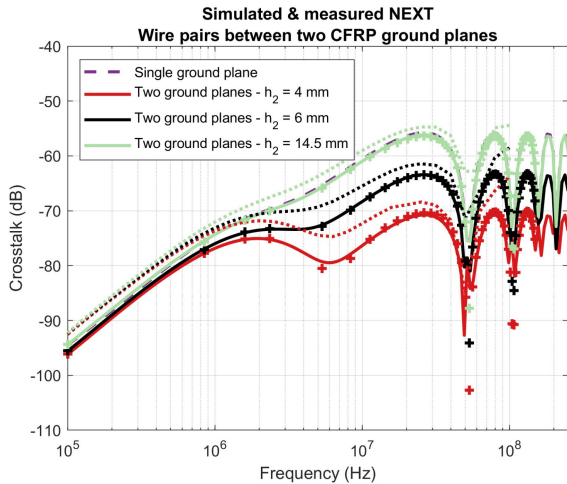


Fig. 13. MTL-DG (solid lines), Feko (markers) and measured (dotted lines) NEXT for two wire pairs between two CFRP ground planes. The lower ground plane is at 3.5 mm, while the height of the other ground plane is changed. MTL-DG result for single CFRP ground plane is shown as reference (purple dashed line).

by uncertainties in the measurement setup, or differences between CFRP properties in simulations and measurements. The variations of the height of the second ground plane show that, similar to the case of a single ground plane [2], the effects of the ground plane to the crosstalk behavior become less when the ground plane is further away. As reference, the simulated result for a single ground plane is also shown in Figs. 12 and 13. For both aluminum and CFRP ground planes, this result coincides with the levels of crosstalk obtained for $h_2 = 14.5$ mm. This confirms that at this distance the second ground plane has no effect anymore. Bringing the second CFRP ground plane closer has little effect on the low-frequency behavior, but it does significantly affect the high-frequency crosstalk. Thus, integrating for instance flexPCBs into thermoplastic material that has conducting fibers, such as CFRP, can greatly reduce high-frequency crosstalk levels.

IV. CONCLUSION

Replacement of conducting materials in aircraft by less conducting composite materials such as CFRP highlights the need for simulation methods that can quickly assess the effects of lossy ground planes to crosstalk behavior. This article presents two methods to incorporate the effects of ground planes with finite conductivity and thickness into MTL models. The first method makes use of a discretization of the ground plane by cylindrical conductors, which have a diameter equal to the thickness of the ground plane. Well-known analytical expressions can still be used for the inductance and capacitance matrices. When the resistance matrix is adapted to include the analytical resistance of cylindrical wires it is found that the resulting MTL-DG crosstalk simulations coincide very well to measured crosstalk and full-wave Feko simulations, for both metallic ground planes as well as CFRP ground planes.

The second method makes use of a ground impedance matrix. Formulas that are used throughout literature to evaluate the effects of lossy ground to overhead transmission lines are adapted to the case of a CFRP ground plane with finite thickness. Computed crosstalk levels obtained with this MTL-GI method coincide very well with the MTL-DG method, as well as with measured crosstalk and full-wave simulations. Both presented MTL methods show the frequency dependent behavior of CFRP ground planes that has been observed in an earlier paper [16]. For low frequencies, EM fields penetrate uniformly through the CFRP ground plane, return currents distributed broadly under the wires, and crosstalk levels are similar to that between wire pairs in free space. For high frequencies, the fields are unable to penetrate the CFRP further than one skin depth. Therefore, return currents concentrate more at the surface and below the current-driving wire and consequently crosstalk levels are equal to those for aluminum ground planes. Both thickness and conductivity of the ground plane are important parameters in the switch in frequency behavior, which appears at the frequency for which the skin depth of the material is in the same order as the thickness of the ground plane.

The MTL-DG method is applied to the case of two wire pairs between two ground planes. Again, for both aluminum and CFRP ground planes the results of calculations and measurements coincide very well. Crosstalk levels indicate that the presence of a second ground plane close to the wire pairs further reduces the crosstalk, when compared to wire pairs in free space and above a single ground plane. For CFRP ground planes this effect is again only observed for the higher frequencies. For low frequencies, the crosstalk is more similar to that of wire pairs in free space. Moving one of the two ground planes away from the wire pairs increases high-frequency crosstalk, up to a point where the levels are equal to the case of a single ground plane.

Concluding, the two MTL methods that include lossy grounds presented in this article yield simulation results that show a perfect match to full-wave simulations and match measurements well. However, computation times on a simple laptop can be from 900 up to 70 000 times better compared to Feko on a 10 core Linux server.

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REFERENCES

- [1] C. R. Paul, *Analysis of Multiconductor Transmission Lines*. New York, NY, USA: Wiley, 1994.
- [2] J. Lansink Rotgerink, H. Schippers, and F. Leferink, "Low-frequency analysis of multiconductor transmission lines for crosstalk design rules," *IEEE Trans. Electromagn. Compat.*, vol. 61, no. 5, pp. 1612–1620, Oct. 2019.
- [3] J. Lansink Rotgerink and H. Schippers, "Crosstalk modelling of unshielded wire pairs," in *Proc. Int. Symp. Electromagn. Compat.*, Gothenburg, Sweden, 2014, pp. 641–646.
- [4] J. Hale, "Boeing 787 from the ground up," *Boeing AERO*, Q04, 2006. [Online]. Available: https://www.boeing.com/commercial/aeromagazine/articles/qtr_4_06/article_04_1.html

- [5] Airbus, "Taking the lead: A350XWB," *Airbus*, 2006. [Online]. Available: https://www.airbusgroup.com/dam/assets/airbusgroup/int/en/investor-relations/documents/2006/untitled/further_pre_a350_xwb_launch_2006.pdf
- [6] J. R. Carson, "Wave propagation in overhead wires with ground return," *Bell Syst. Tech. J.*, vol. 5, no. 4, pp. 539–554, Oct. 1926.
- [7] J. R. Wait, "Theory of wave propagation along a thin wire parallel to an interface," *Radio Sci.*, vol. 7, no. 6, pp. 675–679, Jun. 1972.
- [8] E. D. Sunde, *Earth Conduction Effects in Transmission Lines*, New York, NY, USA: Dover, 1968.
- [9] L. M. Wedepohl and R. Wasley, "Wave propagation in multiconductor overhead lines – Calculation of series impedance for multilayer earth," *IEEE Proc. Inst. Elect. Eng.*, vol. 113, no. 4, pp. 627–632, Apr. 1966.
- [10] F. Rachidi, S. L. Loyka, C. A. Nucci, and M. Ianoz, "A new expression for the ground transient resistance matrix elements of multiconductor overhead transmission lines," *Electr. Power Syst. Res.*, vol. 65, no. 1, pp. 41–46, 2003.
- [11] M. D'Amore and M. S. Sarto, "A new formulation of lossy ground return parameters for transient analysis of multiconductor dissipative lines," *IEEE Trans. Power Del.*, vol. 12, no. 1, pp. 303–314, Jan. 1997.
- [12] F. Rachidi, C. A. Nucci, and M. Ianoz, "Transient analysis of multiconductor lines above a lossy ground," *IEEE Trans. Power Del.*, vol. 14, no. 1, pp. 294–302, Jan. 1999.
- [13] C. R. Paul, "Solution of the transmission-line equations for lossy conductors and imperfect earth," *Proc. IEE*, vol. 122, no. 2, pp. 177–182, Feb. 1975.
- [14] F. Rachidi, "A review of field-to-transmission line coupling models with special emphasis to lightning-induced voltages on overhead lines," *IEEE Trans. Electromagn. Compat.*, vol. 54, no. 4, pp. 898–911, Aug. 2012.
- [15] C. R. Paul, "A brief history of work in transmission lines for EMC applications," *IEEE Trans. Electromagn. Compat.*, vol. 49, no. 2, pp. 237–252, May 2007.
- [16] J. Lansink Rotgerink, F. Happ, and J. van Es, "Crosstalk between wire pairs above a composite ground plane," in *Proc. Int. Symp. Electromagn. Compat.*, Wroclaw, Poland, 2016, pp. 89–93.
- [17] T. Demeester and D. De Zutter, "Quasi-TM transmission line parameters of coupled lossy lines based on the Dirichlet to Neumann boundary operator," *IEEE Trans. Microw. Techn.*, vol. 56, no. 7, pp. 1649–1660, Jul. 2008.
- [18] M. Kechicheb, E. Clavel, and M. R. Mekideche, "Electromagnetic modeling of wire-ground structures by using a MTL based approach," *Automatika*, vol. 57, no. 4, pp. 1035–1044, 2016.
- [19] A. Feko, Feb. 5, 2020. [Online]. Available: <http://www.altairhyperworks.com/product/feko>
- [20] N. Mora *et al.*, "On the validity limits of the transmission line theory in evaluating differential-mode signals along a two-wire line above a ground plane," in *Proc. IEEE Int. Symp. Electromagn. Compat.*, Dresden, Germany, 2015, pp. 797–800.
- [21] F. Happ, F. Gronwald, and H. D. Brüns, "An extension of schelkunoff's shielding theory to anisotropic conducting multilayer materials," in *Proc. IEEE Int. Symp. Electromagn. Compat.*, Dresden, Germany, 2015, pp. 1542–1547.
- [22] C.L. Holloway, M. S. Sarto, and M. Johansson, "Analyzing carbon-fiber composite materials with equivalent-layer models," *IEEE Trans. Electromagn. Compat.*, vol. 47, no. 4, pp. 833–844, Nov. 2005.
- [23] C. L. Holloway, E. F. Kuester, A. E. Ruehli, and G. Antonini, "Partial and internal inductance: Two of clayton r. Paul's many passions," *IEEE Trans. Electromagn. Compat.*, vol. 55, no. 4, pp. 600–613, Aug. 2013.



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