Mutual coupling between antennas in a periodic network using the advanced transverse wave approach for wireless applications

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ABSTRACT

Mutual coupling studies between printed antennas had known some difficulties arising out of either physical phenomena or mainly the computational EM methods used for EM analysis. In this paper, the fast numerical electromagnetic (EM) method so-called Transverse Wave Approach (TWA) and its extended version are presented for EM field modeling of planar structures. Taking into account from benefits of the advanced TWA process, a new efficient technique is introduced to investigate the mutual coupling between antennas in periodic walls. The implementation of this approach with low complexity effort is demonstrated in the context of wireless applications and the obtained simulation results are found to be in good agreement with EM theory and literature.

1. Introduction

From the time when the information and communications technology (ICT) has evolved at astounding rate, the field of radio frequency (RF) and microwave design has pursued firmly behind.

Electromagnetic (EM) simulation has confirmed its existence in this domain since more than two decades and offered for the simulators an important library integrating the most complex and sensitive geometry elements.

Many scientific theses, publications and numerous researches have been devoted to investigate and discuss the notion of mutual coupling between printed antennas for emission or reception (Hui, 2004a; Anselmi et al., 2015; Kalialakis et al., 2015; Ghosh et al., 2016).

Also, theoretical research has shown that the radiation properties of printed antennas (Bernety and Yakovlev, 2016) and mutual coupling between printed dipoles (Khattak et al., 2016) are affected by the radiated fields and the type and number of surface wave modes propagating within the antenna substrate. Both experimental and theoretical researches (Bao et al., 2014; Tsai et al., 2016) have explored the effect of antenna orientation, geometry and substrate on mutual coupling for the H plane and E plane. The antennas in all these studies were linear polarized and radiated the same spatial field (maximum directed broadside to the antenna).

For most applications in communication systems when the scheme of twin antennas is used, the transmitting and receiving antenna array will be closely placed side by side. If the mutual coupling is too strong, the transmitting energy will blockade the receiver; therefore, the reduction of mutual coupling between two arrays is very important.

Thus, many methods have been developed to depress the mutual coupling between antenna elements. Indeed, an efficient technique for decreasing the coupling is manufacturing the isolate below the patch, in order that there is free space below the patch (Ghosh et al., 2016). The horizontal radiation and coupling can be reduced by employing the shorting pins to nullify the capacitive polarization currents of the substrate (Singh, 2018). Some techniques are explicitly designed to suppress the surface wave. It also includes optimizing the antenna dimensions so that the surface wave is not excited (Li et al., 2018), grooving the dielectric, covering the patch by additional dielectric layers (Gharat and Narwade, 2016), or making the dielectric be a band-gap structure by printing various patterns on it (Emadeddin et al., 2017). A

straightforward way to reduce the mutual coupling of monopole antennas on high-impedance ground plane was presented in Lee et al. (2015). A thin piece of conducting tape is placed in the middle between the two horizontal monopoles, extending via the ground plane. The similar idea has been adopted in a MIMO microstrip antenna array to diminish the mutual coupling (Kalialakis et al., 2015; Tu et al., 2017). In order to reduce the mutual coupling between closely-packed antennas elements, a slit pattern etched onto a single ground plane was introduced (Wei et al., 2016).

Ordinarily, we define interaction between two radiating elements by a matrix such as an impedance matrix Z (or admittance matrix). Interaction between these antennas decreases with the term ZI2 (orZ21) when the spacing increases.

For emission, the passive network allows us to define coupling with four complex terms. Reception is another issue in light of the fact that the system turns into a functioning one, as in there is some inside sources forced by the incident wave. It appears to be very obvious that cooperation between the distinctive ports of the system depends on terms identifying with the passive quadripole (Z matrix, for example), and terms relating to the internal sources (Krapez and Dohou, 2014).

The theoretical calculus of mutual coupling between passive or active dipole antennas for emission and reception are given in detail in Daniel (1974). Also, a new definition of mutual impedance for two dipole antennas is introduced in Hui (2004b) and Hor and Janpugdee (2016) in order to characterize the mutual coupling effect between two dipole antennas in a more accurate manner.

All of these studies we mentioned above give an overview either on computation or reduction of the mutual coupling factor between antennas. We propose, in this research, to develop a new approach based on advanced TWA able to diminish, with an efficient way, the mutual coupling between antenna elements used in wireless applications. This can not only enhance the quality of transmitted/emitted information between users but also offer the possibility to our societies to save money by using this new technique instead of other expensive technologies mentioned above.

2. Theoretical background

The present section sets out to present a brief theoretical background of the transvers wave approach TWA and its extended version TWA+. These methods are well developed by our research team in Ayari et al. (2008, 2009a).

2.1. TWA process

Overall, this fast approach is based on the iterative process built from the mutual coupling between incident (A) and reflected (B) waves as shown in the following scheme:

$$\begin{cases} A = \hat{\mathbf{\Gamma}}B \\ A = \hat{\mathbf{S}}B + B^{(0)} \end{cases} \tag{1}$$

Where, $\hat{\Gamma}$ represents the reflection operator connecting the incident and reflected waves in modal domain; $\hat{S}$: designates the diffraction operator linking the incident and reflected waves in spatial domain; $B^{(0)}$: stands for the global excitation wave on the source.

Fig. 1 describes the iterative process of TWA. Indeed, the incident waves are expressed in the spatial domain so that the suitable boundary conditions of EM fields must be fulfilled; this formulation depends on the excitation source. The reflected waves are, nonetheless, formulated in the modal domain considering the properties of EM wave propagating in homogeneous media.

Initially, we assume that the studied structure is excited, for instance, by a bilateral source polarized in x-direction generating waves on both sides from the discontinuity surface Ω. These waves are reflected from both sides i.e., upper and lower medium of the box so as to generate, at first iteration, the incident waves A1(1) and A2(1) which diffract sequentially on the obstacle to produce new reflected waves B1(1) and B2(1). These waves will be reflected again by the upper and lower halves to obtain incident waves in second iteration A1(2) and A2(2) and so on till the instant of system's convergence.
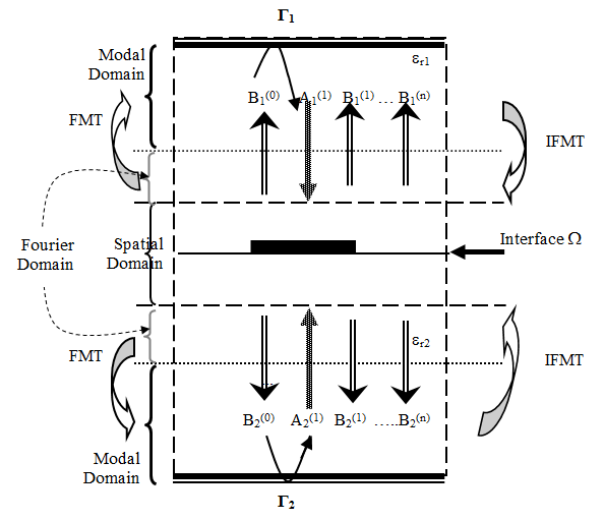

Fig. 1: TWA iterative process

Mathematically, this process is conducted by the fast pixel-mode transform that is performed via two different transformations: the first is between spatial and spectral domains, the second between spectral and modal domains.

The first one is ensured by Fast Fourier transform (FFT) and its inverse IFFT– FFT algorithm is applied to accelerate the transition between these domains. By the way, the Second transformation is guaranteed through Fourier Modal Transform (FMT) and its inverse IFMT (Inverse Fourier-Modal Transform). All of these transformations are shown in Fig. 2.
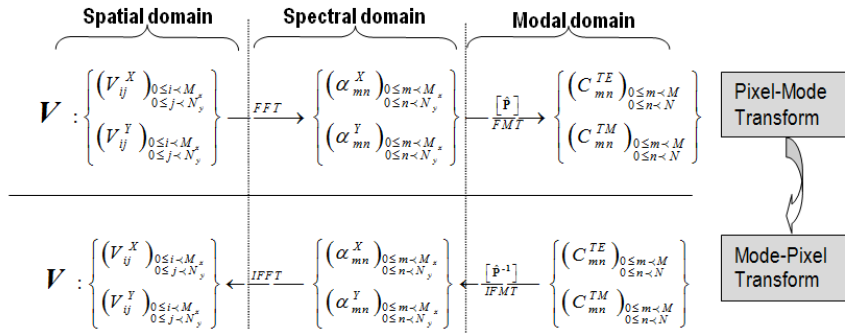
Fig. 2: General principle of the mode-pixel and pixel-mode transforms

More details related to TWA can be found in our work presented in Ayari et al. (2008, 2009b).

2.2. Advanced TWA process

The extended version of the TWA approach offers the possibility to handle input data with non-equispaced discretization. The following scheme summarizes the different transformations from spatial domain to modal domain of a given vector V defined on non-uniform discretization and gives a panorama upon TWA process in its novel version (Fig. 3).
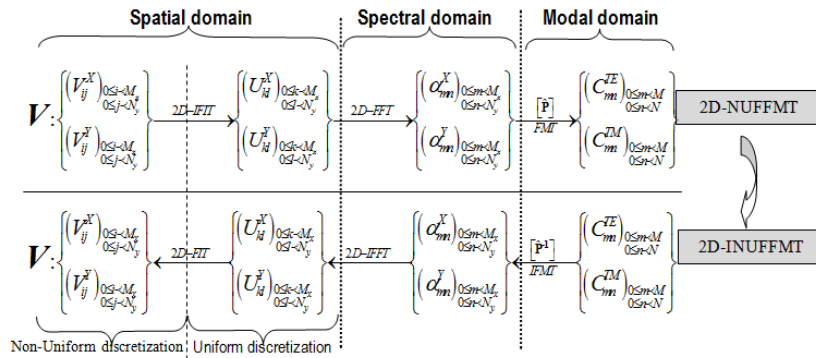

Fig. 3: General prospect on 2D-NUFFMT and its inverse

It is clear that the two-dimensional fast interpolation transforms (2D-FIT) and its inverse has been implemented in the advanced method. We can refer to our work presented in Ayari et al. (2009b) in order to obtain more details upon this advanced approach.

3. Proposed technique and simulation results

The aim of this section is to investigate the mutual coupling between dipole antennas by introducing a simple and efficient method working without any mathematical computation effort. Therefore, we consider two identical dipole antennas already employed having as geometric and modeling simulation parameters Table. The kind of these antennas is used in many wireless applications (Lamminen et al., 2017; Ta et al., 2017; Dadgarpour et al., 2017).

Let $J_p{}^1$ and $E_p{}^1$ be respectively the current density and electric field on excitation domain S^1 situated on the exciting dipole antenna D_1; p refers to the polarization direction of excitation source ($P=X$ our case) and S^1 is nothing but the excitation source sub-domain.

The current density $J_p{}^2$ represents the magnetic field induced on dipole D_2 especially on S^2 which is equal to S^1 in dimension and position with regards to associated dipole (Fig. 4).

We define the coupling parameter between dipoles as the ratio between the current density value on dipole D_2 and the one on the exciting dipole antenna D_1 as the following manner:

$$CJ_P = \left| \frac{J_P^{(\mathbf{D_2})}}{J_P^{(\mathbf{D_1})}} \right| \tag{2}$$

where $\left(J_P^{(\mathbf{D_i})} \right)_{i=1,2}$ characterizes the current density on S^i.

The TWA method offers the possibility to define these terms on the discontinuity plane by the average value of all magnetic field $J_P^{(\mathbf{D_i})}(i_0, j_0)$ on each pixel (i_0, j_0) describing the domain S^i. Thus, $\left(J_P^{(\mathbf{D_i})} \right)_{i=1,2}$ can be calculated as follows:

$$\left(J_P^{(\mathbf{D_i})} \right)_{i=1,2} = \frac{1}{N_S} \sum_{(i_0, j_0) \in S^i} J_P^{(\mathbf{D_i})}(i_0, j_0) \tag{3}$$

where N_S denotes the common total number of pixels defining the domain S^i since all S^i have the same shape.

Table 1: Modeling and geometric constraints

Geometric constraints
Case dimensions: a=b=30mm
Meshing: 32x32
Length of dipole: L=24/32 pixels
Width of dipole: W=1/32 pixels
High of substrate: H=1.52mm

| Modeling constraints |
|---|
| Bilateral source polarized according to x |
| Periodic walls |
| Iteration number: 500 |
| Waveband: Fmin= 1GHz: minimal frequency |
| Fmax= 30GHz : maximal frequency |
| Step_Frq=0.2GHz : frequency step |
| Surface impedance value: Zs=0 (surface impedance sub-domain is equivalent to metallic sub-domain) |
| Permittivity of substrate: 9.8 |
| ε_{r1} =1; ε_{r2} =4.32: permittivity of two sides |
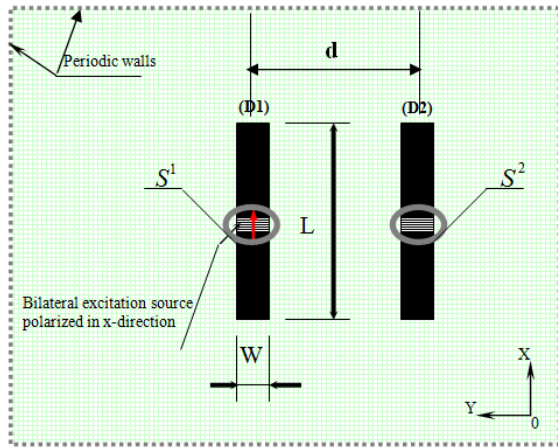
Fig. 4: Two identical dipole antennas studied in periodic network

In reality, the coupling parameter alone cannot characterize completely the coupling between exciting dipole D_1 and dipole D_2 in view of the presence of other factors which can reduce or increase this mutual coupling as mentioned previously. However, the parameter CJ_P can give a quantitative image on it.

We propose to study the effect on this coupling parameter $CJ_{P=X}$ as function of the distance separating the two considered dipole antennas as already depicted in Fig. 4 from excitation source.

The figure depicted in Fig. 5 presents the evolution of the mutual coupling parameter $CJ_{P=X}$ as function of distance d separating dipole D_2 to exciting dipole D_1 at 4 GHz. It shows that d and $CJ_{P=X}$ are inversely proportional, so the coupling decreases when the distance between dipole antennas increases.
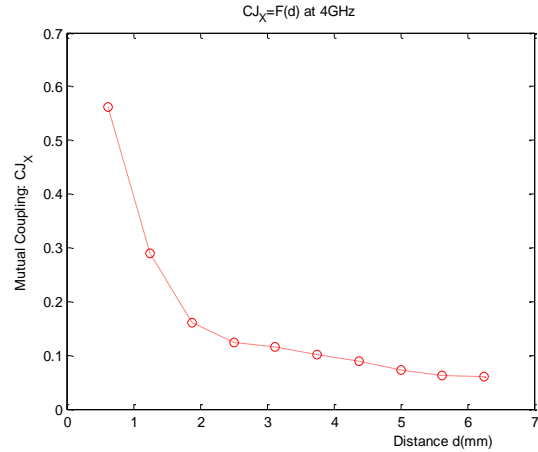

Fig. 5: Evolution of coupling parameter as function of distance **d** at 4 GHz

The distribution of the current density at 4 GHz of the two considered dipoles spaced distance d apart are given for three different values of d in Fig. 6.

These results prove well the decrease of coupling when the distance between dipoles becomes more important.
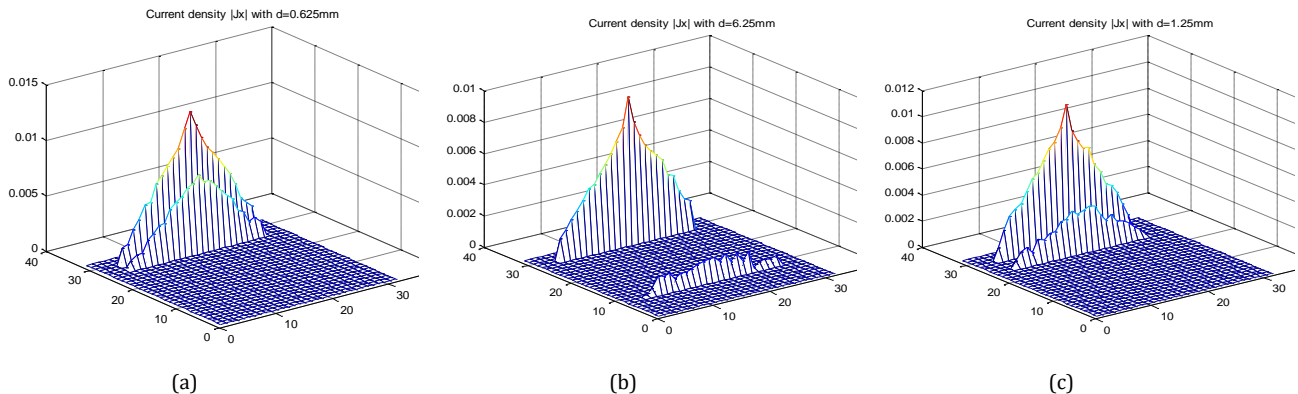

(a)

(b)

(c)

Fig. 6: Distribution of current density at 4 GHz with three different distances (a) d=0.625mm, (b)d=1.25mm and (c)d=6.25mm

Now, let us study the phenomena in presence of N identical dipole antennas D_1, D_2, ..., D_N having L in length and W in width located in box bordered of periodic walls as shown in Fig. 6.

These antennas have been selected as mentioned above to provide a formal framework for EM analysis of wireless applications.

The presence of periodic walls –already implemented in the advanced TWA method– offers the possibility to investigate a huge number of radiant structures and it opens by consequence a good workspace to study an antenna network in the context of wireless applications.

The mutual coupling between exciting dipole D_1 and dipole D_i (i=2, ..., N) can be observed by the distribution of current density of N dipoles at working frequency.

We suppose N=15 and d=0.625mm, the current density $|J_x|$ of the N equispaced (i.e., we assume that two successive dipoles D_i and D_{i+1} are spaced of distance *d*) and identical dipole antennas as depicted above at 4GHz taking into account the simulation parameters already given in Table 1 can be shown as follows:

Let $J_x^{D_k}$, $k \neq 1$ be the current densities characterizing the induced magnetic field on dipoles D_k, $k \neq 1$ (Fig. 7). The coupling parameter CJ_x^k between exciting dipole D_1 and dipole D_k can be defined therefore as follows:

$$CJ_x^k = \left| \frac{J_x^{D_k}}{J_x^{D_1}} \right| ; k \neq 1 \qquad (4)$$
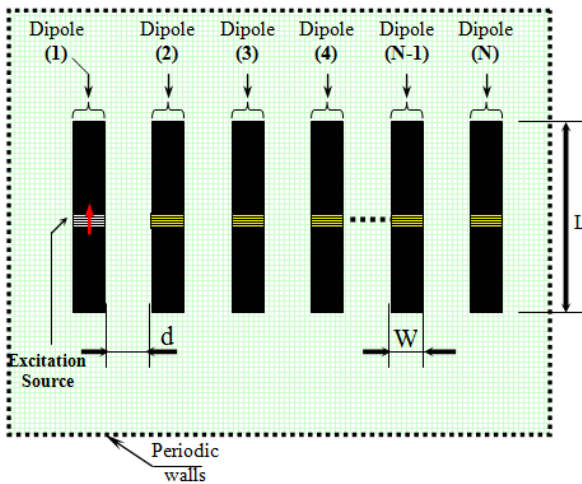

Fig. 7: N dipole antennas studied in periodic network

It is clear that the simulation result given in Fig. 8 is in good agreement with EM theory; but, the impact of the mutual coupling between neighbouring strip dipole antennas can be straightforwardly identified from this distribution.

Indeed, the distribution as shown in Fig. 8 exhibits the influence of the existence of dipoles $(D_i)_{i \neq 1,k}$ on the coupling parameter CJ_x^k in view of the absence of diminishing in the magnitude $|J_x^{D_k}|$ since the condition $\{CJ_x^k < CJ_x^l\}_{k>l}$ has not been completely fulfilled.

Therefore, we should introduce other parameter which allows defining completely the mutual coupling effect between antenna arrays. We talk about mutual impedance in array of N dipole antennas.

To compute the mutual couplings, it suffices to determine the multipole impedance matrix $(Z_{ij})_{1 \leq i,j \leq N}$ defined by Eq. 5.

The Coefficient Z_{jj} is called self-impedance and determines relation between the current and the voltage at the input of antenna elements out of the array and equals to the radiation impedance of a separated j-th antenna element in free space.
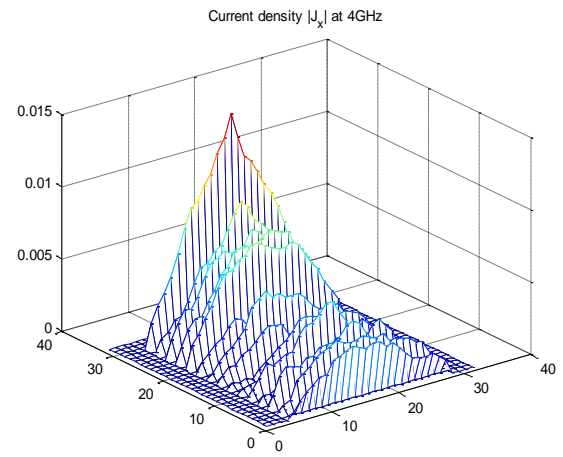

Fig. 8: Distribution of current density at 4 GHz of dipole antenna array

$$
\begin{vmatrix}
E_x^1 \\
E_x^2 \\
\cdot \\
\cdot \\
\cdot \\
E_x^{N-1} \\
E_x^N
\end{vmatrix}
=
\underbrace{
\begin{vmatrix}
Z_{11} & Z_{12} & \cdot & \cdot & \cdot & Z_{1N-1} & Z_{1N} \\
Z_{21} & Z_{22} & \cdot & \cdot & \cdot & \cdot & Z_{2N} \\
\cdot & \cdot & \cdot & & & & \cdot \\
\cdot & \cdot & & \cdot & & & \cdot \\
\cdot & \cdot & & & \cdot & & \cdot \\
Z_{N-11} & & & \cdot & \cdot & Z_{N-1N} \\
Z_{N1} & Z_{N2} & \cdot & \cdot & \cdot & Z_{NN-1} & Z_{NN}
\end{vmatrix}
}_{[\dot{Z}]}
\begin{vmatrix}
J_x^1 \\
J_x^2 \\
\cdot \\
\cdot \\
\cdot \\
J_x^{N-1} \\
J_x^N
\end{vmatrix}
\qquad (5)
$$

The diagonal Z_{ij} of matrix $[Z]$ corresponds to the input impedances of different dipoles in presence of their environment. The detailed calculation of [Z] can be found in Reinhold and Pavel (2000).

To investigate the frequential-dependence of elements of the matrix $[Z]$ can give overall view on the EM behavior of antenna arrays.

4. Conclusion

In this paper, the numerical electromagnetic (EM) method TWA and its extended version TWA+ have been effectively presented to investigate a full-wave of planar antennas. Due to the benefits of this EM approach, a potential and efficient technique working without any mathematical computation effort has been successfully introduced and implemented to investigate the mutual coupling between antennas in periodic network. The simulation results (obtained from our EM software developed by our research team (Ayari et al., 2009c)) proves the efficiency of the proposed technique in the context of wireless applications. This work can be extended to study the integrated circuits with high complexity problem introducing the multi-scale and multi-layer notions in the advanced TWA.

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Compliance with ethical standards

Conflict of interest

The authors declare that they have no conflict of interest.

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