# Non-Resonant Slotted Waveguide Antenna Design Method:

## Inclusive Internal and External Electromagnetic Mutual Coupling Between Slots

**By Michal Grabowski**

In this paper a non-resonant slotted waveguide antenna design method is presented.

In this paper a non-resonant slotted waveguide antenna design method is presented. The internal and external mutual coupling between adjacent radiating slots is considered in the described method. In order to confirm the usefulness of this method, a non-resonant waveguide array antenna with longitudinal slots cut in a broad waveguide wall was designed. The correctness of this method was evaluated by comparison of the results obtained in calculations and their equivalents reached in computer simulations with the use of CST MICROWAVE STUDIO®.

### Introduction

Slotted waveguide antennas (see Fig.1) have been the subject of extensive research for several years. These structures are commonly used in microwave transmitter and receiver devices. Such antennas radiate energy from



**Fig. 1 • Non-resonant slotted waveguide antenna with longitudinal slots cut in a broad waveguide wall.**

feeding waveguide to a free space through several slots cut in a broad or narrow wall of a rectangular waveguide. Special interest in such antennas is caused by their planar and compact structure, high power handling and electrical parameters, such as high efficiency, relatively wide frequency band and good return loss, [1][2][3]. Their additional advantage is the ability to combine vertical slotted waveguides as phased array with shaped and electronically switched multi-beam radiation patterns, which enables the observation of many "moving targets" at the same time, [2][3].

Slotted waveguide array antennas can be realized both as resonant and non-resonant according to the wave propagation inside the waveguide (respectively standing or travelling wave) [4][5][6][7]. Among antennas from the non-resonant subgroup it is possible to distinguish antennas that work over and below resonant frequency (respectively distance between adjacent slots smaller or larger than the half waveguide wavelength). Non-resonant slotted waveguide array antennas, which are tipped with a waveguide termination, provide a feature of squint (not occurring for resonant ones). This effect causes an angle deviation of an end fire direction from a normal to the antenna aperture. Additionally, this deviation is dependent upon an operating frequency and is described as:
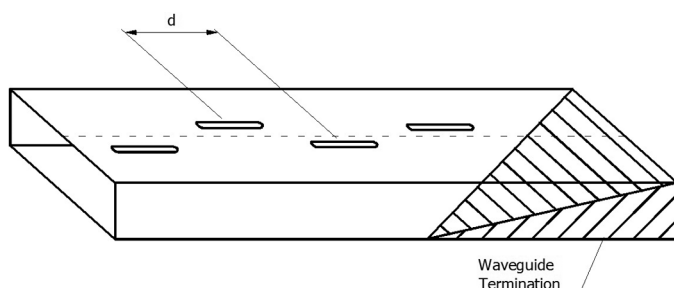
$$\theta_0 = \arcsin\left(\frac{\psi_0 \cdot \lambda}{2\pi d}\right) \quad (1)$$

where $\Psi_0$ is a phase difference of currents feeding adjacent slots, while $d$ is a physical distance between these slots.

One use for these antennas is in a thin array for radio-location devices. It is desirable in these applications to transmit and receive energy to and from a narrow sector of a free space. This requirement is realized by combining several identical parallel rectangular waveguide antennas with close to half-wave slots cut in their walls [3][8][10]. The radiation patterns of such antennas are dependent on complex amplitude distributions of equivalent currents feeding slots (currents corresponding to surface currents superposition which flows over walls of waveguide and crossing slots). Appropriate amplitude distributions of current fed radiating slots are determined during the synthesis process for a definite spatial radiation pattern. Multi-slots array spatial radiation patterns might be determined by several methods, described in literature [2][3], such as: Taylor, Dolph-Tschebyshev or Fourier Integral Methods.

The design of the slotted waveguide array antenna is a fairly complicated task. It requires including an influence of the internal (by a supplying slots waveguide) and the external (through the open space) mutual coupling between radiating slots on a radiation pattern. Such mutual coupling distorts a radiation pattern of an antenna array as a rule, what can be mostly seen as a side lobes level increased, widening of a main lobe and an angle deviation of an end fire direction from a normal to an antenna aperture. These effects are especially significant for array antennas with a small number of slots ($N \leq 12$). In an extreme situation mutual coupling can lead to an antenna physical non-realizability, which was studied in [10].

In this paper a design method of a slotted waveguide antenna with longitudinal slots cut in a broad waveguide wall is proposed. In the presented method the internal and external mutual coupling mentioned above is taken into account. This method based on the Full-Wave method ([6][7][8][10]) is an extension of the external mutual coupling effect described in [9][11]. In order to confirm the usefulness of this method, the non-resonant waveguide array antenna with 12 longitudinal slots cut in a broad waveguide wall was designed. The effect of internal and external mutual coupling was evaluated by the comparison of the radiation pattern obtained with the presented method, and the pattern reached with what is commonly known as the Energetic Method, [3] (not including mutual coupling between slots). The correctness of the proposed method was confirmed by comparing results reached in computer simulations with CST MICROWAVE STUDIO.

## Design Method of Slotted Waveguide Antennas Including Internal and External Mutual Coupling Between Slots

The design of a multi-slots waveguide antenna requires determining the slots' geometric dimensions and their location in a waveguide dependent on the excitation amplitude distribution of equivalent feeding slots. Such antennas can be analyzed by various methods. Unfortunately the simplest methods such as the Energetic ([3]) and the Full-Wave ([6][8][10]) do not take into consideration both internal and external mutual coupling effects between slots. Thus the results are encumbered with errors.

The most significant effects which should be considered in the design process are internal and external mutual coupling between slots. The internal mutual couplings are caused by the partial reflections of the incident electromagnetic wave from succeeding slots in a waveguide. These partial reflections cause a considerable displacement of the EM field inside the waveguide. The external mutual couplings are induced as the partial wave pervades from one slot to the others through the open space. Similar to internal couplings, these pervasions cause the EM field displacement inside the waveguide and lead to the change of the slots' immittance. The last significant effect which should be taken into account in the waveguide antenna design process is to consider the reactance or susceptance properties of slots, when their lengths are different from that of resonance.

In non-resonant antennas distance between adjacent slots $d$ is different then $\lambda_{01}/2$, where $\lambda_{01}$ is a waveguide
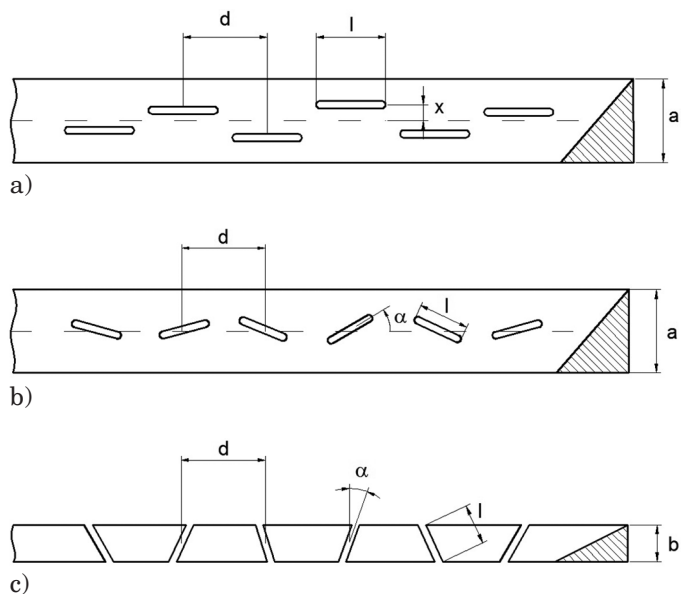


a)

b)

c)

**Figure 2 •** Basic solutions of the slotted waveguide antennas with: (a) longitudinal slots, (b) inclined slots cut in a broad wall, (c) inclined slots cut in a narrow wall.

wavelength. Such antennas are tipped with waveguide terminations, resulting in a wave propagating in the waveguide being close to the travelling wave. In Fig.2 are shown the three basic solutions of slotted waveguide antennas. Respectively, there are antennas with longitudinal slots cut in a broad wall (Fig. 2a), and inclined slots cut in a broad (Fig. 2b) and a narrow wall (Fig. 2c) of the rectangular waveguide.

In order to reach an optimal coupling of slots and feeding waveguide, the length of all slots near to the resonant length are desired (typically it is $l \epsilon \langle 0.9 \cdot l_r, 1.1\ l_r \rangle$). By the resonance frequency we mean the frequency when an imaginary part of impedance or admittance represented the slot equals zero. So for resonance slots their impedance or admittance are real values.

As mentioned above the design method proposed in this article combines the Full-Wave Method including only internal mutual coupling and the method presented in [9][11] including only external mutual coupling. The presented method extended by a correction stage allows us to take into account the influence of slots' susceptance or reactance on the radiation pattern, when their length are different than resonant ($l \neq l_r$). The basis of this method is to evaluate partial reflections of the electromagnetic wave from succeeding slots which cause a considerable displacement of the EM field inside the waveguide. An essential challenge with this approach is to calculate a complex amplitude distribution of waves propagating from the waveguide input to the matched termination and in the reverse direction (see Fig. 3). Depending on slot types, such structures can be analyzed as different equivalent circuits. For inclined slots cut in a broad wall of rectangular waveguide (see Fig. 2b) each of these slots become the serial lumped impedance. On the contrary longitudinal shunt slots cut in a broad wall (see Fig. 2a) and inclined slots cut across a narrow wall (see Fig. 2c) can be determined as parallel admittance inserted in a waveguide.
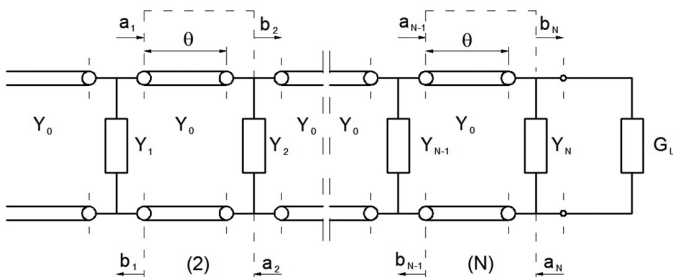


Figure 3 • Electrical schemes of the *N*-slotted waveguide antennas equivalent circuits, with slots represented by normalized parallel admittance.
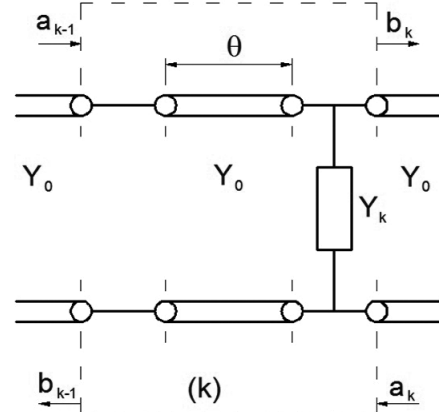


Figure 4 • Electrical scheme of the two-port substitute circuit where the slot is represented by parallel admittance.

In the next part of this article the algorithm of our design method is limited only for waveguide antennas with longitudinal slots cut in a broad waveguide wall. The equivalent circuit of such an *N*-slotted waveguide antenna, where each slot is represented by a normalized parallel admittance ($y_k = Y_k/Y_0 = g_k+jb_k$) inserted into a long transmission line with a characteristic admittance $Y_0$, is shown in Fig. 3.

The circuit shown in Fig. 3 can be analyzed as a cascade connection of *N* two-port substitute circuits shown in Fig. 4.

## Algorithm of the Non-Resonant Slotted Waveguide Antenna Design Method

As mentioned above, the proposed algorithm enables us to design antennas including the internal and external mutual coupling between slots. At first for the desirable complex amplitude distribution of the feeding slots equivalent current $I_k$ and desirable value of the loss coefficient $\chi_N$ (defined as the ratio of the power lost in the waveguide termination to the input power) it is necessary to recalculate the relative power distribution radiated by each slot $P_k$ normalized to the antenna input power $P_0$.

In the next stage the conductance of all antenna slots must be evaluated. In this stage only internal mutual coupling is considered and all slot lengths are assumed as the resonant. The Full-Wave Method presented in [6][8][10] will be used.

The transmission matrix T of the circuit shown in Fig. 4 equals the product of the transmission matrixes of a long transmission line with a characteristic admittance $Y_0$ and a parallel inserted normalized admittance $y_k$. The transmission matrix of the circuit, shown in Fig.4, equals ([6][8][10]):

$$\begin{bmatrix} T_{11}^{(k)} & T_{12}^{(k)} \\ T_{21}^{(k)} & T_{22}^{(k)} \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \cos\theta - j\sin\theta & 0 \\ 0 & \cos\theta + j\sin\theta \end{bmatrix} \cdot \begin{bmatrix} 2-g_k-jb_k & -g_k-jb_k \\ g_k+jb_k & 2+g_k+jb_k \end{bmatrix} \quad (2)$$

After multiplication Equation (2) equals:

$$T_{11}^{(k)} = \left(1 - \frac{g_k}{2}\right)\cos(\theta) - \frac{b_k}{2}\sin(\theta) - j\left[\left(1 - \frac{g_k}{2}\right)\sin(\theta) + \frac{b_k}{2}\cos(\theta)\right]$$

$$T_{12}^{(k)} = -\frac{g_k}{2}\cos(\theta) - \frac{b_k}{2}\sin(\theta) + j\left[\frac{g_k}{2}\sin(\theta) - \frac{b_k}{2}\cos(\theta)\right]$$

$$T_{21}^{(k)} = \frac{g_k}{2}\cos(\theta) - \frac{b_k}{2}\sin(\theta) + j\left[\frac{g_k}{2}\sin(\theta) + \frac{b_k}{2}\cos(\theta)\right]$$

$$T_{22}^{(k)} = \left(1 + \frac{g_k}{2}\right)\cos(\theta) - \frac{b_k}{2}\sin(\theta) + j\left[\left(1 + \frac{g_k}{2}\right)\sin(\theta) + \frac{b_k}{2}\cos(\theta)\right]$$

$$(3)$$

According to the Fig.4 all incident and reflected waves at the input of the $k$-th circuit ($a_{k-1}$ and $b_{k-1}$) are related to the appropriate waves at the output of this circuit ($a_k$ and $b_k$) through the following simple formula:

$$\begin{bmatrix} b_{k-1} \\ a_{k-1} \end{bmatrix} = \begin{bmatrix} T_{11}^{(k)} & T_{12}^{(k)} \\ T_{21}^{(k)} & T_{22}^{(k)} \end{bmatrix} \cdot \begin{bmatrix} a_k \\ b_k \end{bmatrix} \quad (4)$$

So, complex amplitudes at the input of the $k$-th circuit equal:

$$b_{k-1} = T_{11}^{(k)} \cdot a_k + T_{12}^{(k)} \cdot b_k$$
$$a_{k-1} = T_{21}^{(k)} \cdot a_k + T_{22}^{(k)} \cdot b_k \quad (5)$$

For $a_{k-1} = A_{k-1} + i B_{k-1}$, $b_{k-1} = C_{k-1} + i D_{k-1}$, $a_{k-1} = A_k + i B_k$ and $b_k = C_k + i D_k$, Equation (5) is given as follows ([6][8][10]):

$$C_{k-1} + i \cdot D_{k-1} = T_{11}^{(k)}\left(A_k + i \cdot B_k\right) + T_{12}^{(k)}\left(C_k + i \cdot D_k\right)$$
$$A_{k-1} + i \cdot B_{k-1} = T_{21}^{(k)}\left(A_k + i \cdot B_k\right) + T_{22}^{(k)}\left(C_k + i \cdot D_k\right) \quad (6)$$

According to Equations (4) and (6) the parameters $A_{k-1}$, $B_{k-1}$, $C_{k-1}$ and $D_{k-1}$ take the following forms:

$$A_{k-1} = \left(1 + \frac{g_k}{2}\right)\left[C_k\cos(\theta) - D_k\sin(\theta)\right] + \frac{g_k}{2}\left[A_k\cos(\theta) - B_k\sin(\theta)\right] +$$
$$- \frac{b_k}{2}\left[A_k\sin(\theta) + B_k\cos(\theta) + C_k\sin(\theta) + D_k\cos(\theta)\right]$$

$$B_{k-1} = \left(1 + \frac{g_k}{2}\right)\left[C_k\sin(\theta) + D_k\cos(\theta)\right] + \frac{g_k}{2}\left[A_k\sin(\theta) + B_k\cos(\theta)\right] +$$
$$+ \frac{b_k}{2}\left[A_k\cos(\theta) - B_k\sin(\theta) + C_k\cos(\theta) - D_k\sin(\theta)\right]$$

$$(7)$$

$$C_{k-1} = \left(1 - \frac{g_k}{2}\right)\left[A_k\cos(\theta) + B_k\sin(\theta)\right] - \frac{g_k}{2}\left[C_k\cos(\theta) + D_k\sin(\theta)\right] +$$
$$- \frac{b_k}{2}\left[A_k\sin(\theta) - B_k\cos(\theta) + C_k\sin(\theta) - D_k\cos(\theta)\right]$$

$$D_{k-1} = \left(1 - \frac{g_k}{2}\right)\left[-A_k\sin(\theta) + B_k\cos(\theta)\right] + \frac{g_k}{2}\left[C_k\sin(\theta) - D_k\cos(\theta)\right] +$$
$$- \frac{b_k}{2}\left[A_k\cos(\theta) + B_k\sin(\theta) + C_k\cos(\theta) + D_k\sin(\theta)\right]$$

For the determined conductance in the first stage of the presented algorithm the power amplitude distribution $P_k$ it is possible to recalculate normalized values of the conductance for all slots. The conductance of the $k$-th slot, is defined as, [6][8][10]:

$$g_k = \frac{P_k}{|a_k + b_k|^2} = \frac{P_k}{\left(A_k + C_k\right)^2 + \left(B_k + D_k\right)^2} \quad (8)$$

where $A_k$, $B_k$, $C_k$ and $D_k$ are determined in Equation (7).

Assuming that the antenna is loaded with the ideally matched waveguide termination ($g_{N+1} = 1$), the design begins from the last $N$-th circuit. Then the absolute square of the amplitude $b_N$ equals the power lost in matched termination ($|b_n|^2 = X_N$) and the amplitude of the wave reflected from the termination equals $a_N = 0$ (see Fig.3). For this assumption the real and imaginary amplitudes of the incident and reflected waves for the waveguide termination equal: $A_N = 0$, $B_N = 0$, $C_N = \sqrt{X_N}$, $D_N = 0$. Conductance of the $N$-th slot (see Fig.4) evaluated with Equation (8) is given as follows ([6][8][10]):

$$g_N = \frac{P_N}{C_N^2} = \frac{P_N}{X_N} \quad (9)$$

For known conductance of the $N$-th slot it is possible to determine (with Equation (7)) real and imaginary parts of incident $a_{N-1}$ and reflected $b_{N-1}$ waves on the input of the $N$-th circuit (see Fig.3). At the beginning it is assumed that slot susceptance in Equation (7) equals zero ($b_k = 0$). For known components $A_{N-1}$, $B_{N-1}$, $C_{N-1}$ and $D_{N-1}$, normalized conductance $g_{N-1}$ can be evaluated with Equation (9). In the analog way it is possible to evaluate the amplitudes of the incident and reflected waves for the $N-1$-st circuit. In a similar way the conductance of all slots cut in the antenna can be evaluated. These values take into consideration only internal mutual couplings between slots.

In the third stage of the presented algorithm, the slots' geometrical parameters which are crucial for the coupling between the slot and the feeding waveguide (such as the displacement of a slot from axis of a waveguide $x_k$ and the length of the slot $l_k$) should be evaluated. For known slot conductance values (evaluated in the previous stage) the geometrical parameters such as $x_k$ and $l_k$ can be obtained with proper closed form formulas which are presented in several publications. The most popular and accurate closed form formulas which allow us to obtain complex admittance for slots cut in the broad waveguide wall were presented by Oliner in [12]. For known slots geometrical parameters such as $x_k$ and $l_k$ it is needed to recalculate the admittance of each slot. Recalculated in such way, admittances are complex values and include the susceptance properties of slots.

In the last stage of the proposed method, for known slot admittance $y_k = g_k + i \cdot b_k$ (evaluated in the previous stage) it is possible to recalculate components $A_k$, $B_k$, $C_k$ and $D_k$ more precisely to include slot susceptance (see Equation (7)). In this stage the external mutual coupling between slots will be taken into account. Similar to the second stage of this algorithm, the design of the antenna begins from the last $N$-th circuit ($A_N = 0$, $B_N = 0$, $C_N = \sqrt{X_N}$, $D_N = 0$). So, the admittance of the $N$-th slot can be evaluated with Equation (9).

In order to evaluate the external mutual coupling between the $N$-th slot and rest of the waveguide slots, the $N$-th slot admittance including the mutual admittance between this slot and rest of the slots should be recalculated. These calculations can be done with formulas presented in [9]. In this article the mutual slot admittance was evaluated based on Babinet's principle linking the performance of the slot array to the equivalent dipole array. The active admittance of the longitudinal slot cut in a broad waveguide wall (consider interrelations among all slots) can be evaluated as follows, [9]:

$$y_k^A = \frac{\dfrac{4 \cdot a \cdot \lambda_{01}}{0.61 \cdot \pi \cdot b \cdot \lambda_0} \cdot \left[\cos\left(\dfrac{\pi \cdot l_k}{\lambda_0}\right) - \cos\left(\dfrac{\pi \cdot l_k}{\lambda_{01}}\right)\right]^2 \cdot \sin^2\left(\dfrac{\pi \cdot x_k}{a}\right)}{\dfrac{Z_k}{73} - \sum_{l=1}^{N} {}' \dfrac{Z_{lk}^2}{73 \cdot Z_l}} \quad (10)$$

where $a$ and $b$ are respectively interior broad and narrow wall dimensions of the rectangular waveguide, $\lambda_0$ is the free space wavelength, $\lambda_{01}$ is the waveguide wavelength, $Z_k$ and $Z_l$ are the self-impedance of the dipoles equivalent via Babinet's principle to the self-admittance of the $k$-th and $l$-th slots, $Z_{kl}$ is the mutual admittance between $k$-th and $l$-th dipoles. The impedance of the $k$-th dipole equivalent via Babinet's principle to the $k$-th longitudinal slot cut in a broad waveguide wall can evaluated with the equation:

$$Z_k = \frac{73}{y_k} \cdot \frac{4 \cdot a \cdot \lambda_{01}}{0.61 \cdot \pi \cdot b \cdot \lambda_0} \cdot \left[\cos\left(\frac{\pi \cdot l_k}{\lambda_0}\right) - \cos\left(\frac{\pi \cdot l_k}{\lambda_{01}}\right)\right]^2 \cdot \sin^2\left(\frac{\pi \cdot x_k}{a}\right) \quad (11)$$

where $y_k$ is the normalized slot self-admittance recalculated in the third stage of the presented algorithm. The mutual impedance $Z_{lk}$ between $k$-th and $l$-th dipoles (see Fig.5) can be evaluated with formulas presented in [13]. The real and imaginary part of this impedance can be evaluated as follows, [13]:

$$R_{lk} = -30 \int_{s=-L_1/2}^{s=L_1/2} \left\{ \frac{1}{\rho^2} \left[ \sin(2\pi r_k) \frac{s_z + z_0 + \frac{l_k}{2}}{r_k} + \sin(2\pi r_l) \frac{s_z + z_0 - \frac{l_k}{2}}{r_l} + \right. \right.$$
$$\left. - 2\cos(\pi L_k)\sin(2\pi r)\frac{s_z + z_0}{r} \right] \cdot \left( s_x^2 + y_0 s_y + s_y^2 \right) +$$
$$\left. + \left[ \left( 2\frac{\sin(2\pi r)\cdot\cos(2\pi r)}{r} - \frac{\sin(2\pi r_k)}{r_k} - \frac{\sin(2\pi r_l)}{r_l} \right) \cdot s_z \right] \right\} \cdot \left\{ \frac{\sin\left[ 2\pi\left(\frac{l_l}{2} - |s|\right)\right]}{s} \right\} ds \quad (12)$$

$$X_{lk} = -30 \int_{s=-L_1/2}^{s=L_1/2} \left\{ \frac{1}{\rho^2} \left[ \cos(2\pi r_k) \frac{s_z + z_0 + \frac{l_k}{2}}{r_k} + \cos(2\pi r_2) \frac{s_z + z_0 - \frac{l_k}{2}}{r_l} + \right. \right.$$
$$\left. - 2\cos(\pi L_k)\cos(2\pi r)\frac{s_z + z_0}{r} \right] \cdot \left( s_x^2 + y_0 s_y + s_y^2 \right) +$$
$$\left. + \left[ \left( 2\frac{\cos(2\pi r)\cdot\cos(2\pi r)}{r} - \frac{\cos(2\pi r_k)}{r_k} - \frac{\cos(2\pi r_l)}{r_l} \right) \cdot s_z \right] \right\} \cdot \left\{ \frac{\sin\left[ 2\pi\left(\frac{l_l}{2} - |s|\right)\right]}{s} \right\} ds \quad (13)$$

where $s_z = s \cdot \cos(\theta)$, $s_y = s \cdot \sin(\theta) \cdot \sin(\phi)$, $s_x = s \cdot \sin(\theta) \cdot \cos(\phi)$, $\rho = \sqrt{s_x^2 + (y_0 + s_y)^2}$, $r = \sqrt{r^2 + (z_0 + s_z)^2}$, $r_k = \sqrt{\rho^2 + \left(z_0 + s_z + \frac{l_k}{2}\right)^2}$, $r_l = \sqrt{\rho^2 + \left(z_0 + s_z - \frac{l_k}{2}\right)^2}$.

The mutual impedance between two dipoles equals: $Z_{lk} = R_{lk} + i \cdot X_{lk}$.

Evaluated with Equation (10), active admittance of the $N$-th slot $y_N^A = g_N^A + i \cdot b_N^A$ can be used to recalculate components of incident and reflected waves $A_{N-1}$, $B_{N-1}$, $C_{N-1}$ and $D_{N-1}$ for $N$-1-st circuit (see Fig.3). These values can be reached with Equation (7), but as the slot conduc-
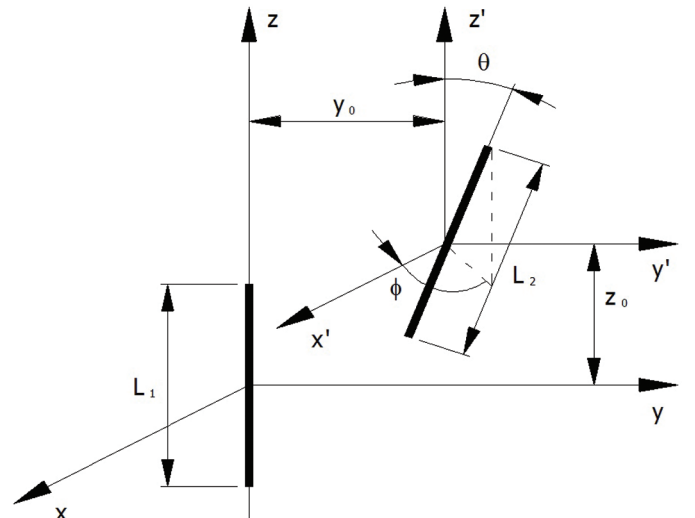


**Figure 5 • Scheme of two independent dipoles geometrical dimensions.**

tance and susceptance $g_k$ and $b_k$, values $b_N^A$ and $g_N^A$ obtained with Equation (10) should be used. In a similar way the conductance (with Equation (8)) and afterwards the active admittance (with Equation (10)) of the *N-1*-st slot can be evaluated. In this manner it is possible to evaluate the amplitudes of the incident and reflected waves and admittance values for all antenna slots.

The phase of the equivalent current flows through the *k*-th slot can be recalculated as follows, ([6][8][10]):

$$\varphi_k = \arg\left(a_k + b_k\right) = arctg\left(\frac{B_k + D_k}{A_k + C_k}\right) + m \cdot \pi \qquad (14)$$

where $m=1,2,3,...$ . Parameter $m$ is chosen in that way to minimize the phase deviation $\phi_k$ from the linear distribution. The phase of the equivalent current obtained with Equation (14) flowing through all slots enables us to determine the deviations of the phase angles from the linear distribution $\delta\phi_k$. This deviation for the *k*-th slot can be given as follows ([6][8][10]):

$$\delta\varphi_k = \left(\frac{2\pi}{\lambda_{01}d} + \pi\right)(N-k) - \varphi_k \qquad (15)$$

**Design of the Non-Resonant Waveguide Array Antenna with 12 Longitudinal Slots Cut in a Broad Waveguide Wall.**

The method proposed in this article was used to design a non-resonant waveguide antenna with 12 longitudinal slots alternating placed on both sides of a rectangular waveguide's broad wall axis (see Fig. 2a). The waveguide dimensions are as follows: the broad and narrow wall dimensions: a = 22.86 mm, b = 10.16 mm, thickness of the

wall t = 1.27 mm (WR90 waveguide standard). The Taylor's amplitude distribution ([2][3]) of the feeding slots equivalent current with parameters α = 1.5 and $n$ = 8 was chosen. Values of desirable current amplitudes are shown in column 2 of Tab. 1. For calculations it is assumed: the operation frequency $f_0$ = 9.35GHz, distance between adjacent slots $d$ = 0.45 $\lambda_{01}$ = 20.27mm, and the loss coefficient $\chi_N$ = 0.15 (efficiency coefficient of the antenna equals η = 85%). Length and width of slots were chosen in that way to obtain resonant slots (slots' susceptance equal zero). For this assumption slots' length and width equal: L = 15.35mm, W = 1mm. In Tab. 1 shows the slots' conductance normalized to the waveguide admittance without mutual coupling consideration (column 3) and with internal and external mutual coupling between slots. In the last column of Tab. 1 the phase deviation from the linear distribution of the feeding slots currents $\delta\phi_k$ is also shown.

Fig. 6 shows the normalized radiation patterns obtained with the proposed design method including internal and external mutual coupling between slots and with the Energetic method [3]. In order to confirm the usefulness of the presented method in Fig.6 the computer simulation results obtained in CST MICROWAVE STUDIO are also presented.

In comparing the radiation patterns shown in Fig. 6 we observe differences between results recalculated with the Energetic Method (mutual coupling not included) and with the method proposed in this article. These differences are caused by not considering the mutual coupling between slots in the Energetic Method. The most significant effects of the internal mutual coupling influence on the radiation pattern are an increase of the first sidelobes

| Slot No. | Current Amplitude Distribution | Slot's conductance without mutual coupling consideration | Slot's conductance with mutual coupling consideration | Deviation of phase angle from linear distribution |
|---|---|---|---|---|
| $k$ | $l_k$ | $g_k$ | $g_k$ | $\delta\phi_k['] $ |
| 1 | 0.30 | 0.0138 | 0.0100 | -6.51° |
| 2 | 0.38 | 0.0232 | 0.0157 | -11.59° |
| 3 | 0.58 | 0.0550 | 0.0374 | -17.12° |
| 4 | 0.77 | 0.1012 | 0.0745 | -22.13° |
| 5 | 0.92 | 0.1596 | 0.1345 | -25.72° |
| 6 | 1.00 | 0.2260 | 0.2265 | -26.91° |
| 7 | 1.00 | 0.2920 | 0.3496 | -24.74° |
| 8 | 0.92 | 0.3465 | 0.4612 | -18.95° |
| 9 | 0.77 | 0.3739 | 0.4622 | -11.31° |
| 10 | 0.58 | 0.3435 | 0.3234 | -5.07° |
| 11 | 0.38 | 0.2261 | 0.1528 | -1.61° |
| 12 | 0.30 | 0.1765 | 0.0955 | 0.00° |

**Table 1 • Electrical parameters of the designed antenna.**

pair level and the distortion of the main lobe. Radiation patterns obtained with the Full-Wave Method and in CST MICROWAVE STUDIO simulations were also compared. For the first pair of sidelobes the biggest difference was observed. It is worthwhile to emphasize that results obtained in computer simulations and reached with the proposed method are closer in the comparison to results evaluated with the Energetic Method. In Fig.7 the characteristics of the antenna return loss and loss coefficient $\chi_N$ (loss in the matched waveguide termination) obtained with CST MICROWAVE STUDIO are presented. The loss coefficient for the operation frequency $f_0$ = 9.35 GHz equals $\chi_N$ = 0.158(−8.015 dB) and is close to the desirable value $\chi_N$ = 0.15. The presented simulation results and results reached with the proposed method are in a good agreement. It allows us to draw the conclusion that the proposed method is correct and can be used in the design process of non-resonant slotted waveguide antennas.

## Conclusions

In this article a design method of non-resonant slotted waveguide array antennas including internal and external mutual coupling between slots was proposed. This method allows us to determine the geometrical dimensions of slots and their location in the waveguide (which are crucial for the coupling between the slot and the feeding waveguide) dependent on the desirable amplitude distributions of equivalent currents feeding slots. In order to confirm its usefulness, the proposed method was used to design a non-resonant waveguide antenna with 12 longitudinal slots alternately placed on both sides of a rectangular waveguide's broad wall axis. As the desirable choice, the Taylor amplitude distribution of the feeding slots equivalent currents was chosen.

The correctness of this method was verified by the results obtained in calculations and their equivalents reached in computer simulations with CST MICROWAVE STUDIO. Both patterns were in the agreement and main differences were for the level of the first side lobes. It is worthwhile to emphasize

that these differences were much smaller than the differences between patterns obtained with the Energetic Method and in the simulations. Strong agreement between results confirms the usefulness of the proposed method.

The influence of the external and internal mutual couplings between slots on the radiation pattern was estimated by the comparison of the radiation patterns reached in the Energetic Method (couplings not included) and the proposed method. The obtained patterns were
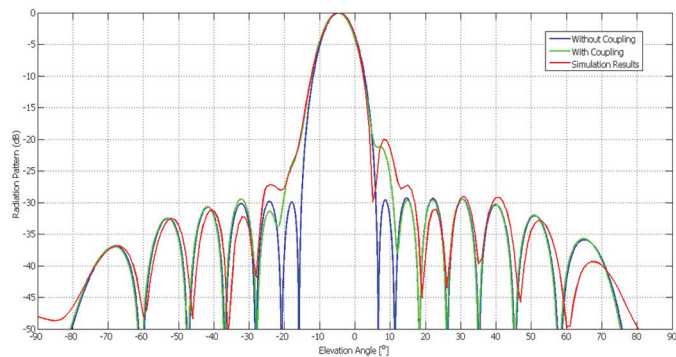
**Figure 6 • The normalized radiation pattern of the designed antenna obtained with the presented method (green line), Energetic Method (blue line) and the CST MICROWAVE STUDIO simulations (red line).**



**Figure 7 • The antenna return loss (red line) and loss coefficient $\chi_N$ characteristics obtained in CST MICROWAVE STUDIO simulations.**

especially different for the side lobe levels and the distortion of the main lobe. Additionally this effect caused an increase of the end fire angle deviation from the normal to the antenna aperture.

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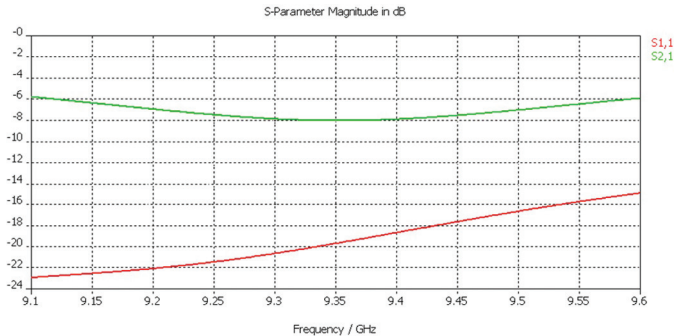
**Note:** CST MICROWAVE STUDIO® is a registered trademark of CST in North America, the European Union, and other countries.

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**About the Author:**

Michal Grabowski is an Antenna Design Engineer with Cobham Antenna Systems (Marlow), The Fourth Avenue, Marlow Buckinghamshire, SL7 1TF, UK. He is a Ph.D. candidate in Electrical Engineering at the Institute of Radioelectronics Warsaw University of Technology, Nowowiejska st. 15/19, 00-665 Warsaw, Poland.