FERRITE LOADED WAVEGUIDES

Ferrites, as well as their technological applications, have been known for some time. *Magnetite*, the first known magnetic material, is actually a ferrous ferrite. In 1269 Peter Peregrinus gave a detailed description of a compass made with a floating *magnetite* needle, probably a Chinese invention. The first experimental ferrite device in microwave technology was demonstrated in 1949 [1]. Since then, applications of artificial ferrite materials in microwave technology have grown rapidly and have become a mature technology, which has been discussed in many classical textbooks [2] - [10]. A good historical survey of the beginnings of microwave ferrite technology can be found in Button [1]. A complete bibliography containing the most relevant contributions in this field during the years that followed can be found in [11] and [12]. Finally, a survey of ferrite technology in Europe, the United States and Japan can be found in [13], [14] and [15], respectively.

This article describes the main physical effects due to the propagation and guidance of electromagnetic waves in ferrite loaded waveguides useful in microwave technology. The linear approach, in which the high frequency magnetic susceptibility of the ferrite is a function of the internal static magnetic field, will be considered valid. This approach includes the analysis of exchange free electromagnetic waves, as well as magnetostatic waves and other approximations, but not the analysis of spin waves and nonlinear effects, magnetoelastic waves and other complex interactions.

The choice of units in the analysis of microwave applications of ferrites presents some particularities. SI (International System) units are preferred for most of the electronic and electrical engineers. Nevertheless, c.g.s. units are mainly used by researchers in the area of the constitutive electromagnetic properties of ferrites. The usage in this text is a compromise between both alternatives. Since in the linear theory the equations for the static biasing magnetic field are decoupled from the radio frequency field equations, we will use c.g.s. units in the derivation of the internal magnetostatic field, as well as in the expression of the magnetic permeability in terms of the static bias field $\mathbf{H}_0$, measured in Oersted, and the saturation magnetization $4\pi\mathbf{M}_0$, with M_0 measured in Gauss ($1\,Oe = 4\pi\,G$). For the electromagnetic r.f. equations we will use SI units.

FUNDAMENTALS OF ELECTRODYNAMICS OF FERRITE MATERIALS

As long as the linear approach remains valid, the problem of finding the radio frequency electromagnetic field inside a magnetized ferrite can be divided into three steps. First we must find the internal static field $\mathbf{H}_0$ as a function of the external applied static field $\mathbf{H}_{ext}$. Then we must obtain the RF magnetic permeability tensor that will be a function of the internal static field. Finally, we must solve the Maxwell equations for the RF field with the appropriate boundary conditions. Notice that, in the linear approach, the equations for the static field $\mathbf{H}_0$ remains independent from the equations for the RF field. The coupling between these two fields occurs only by means of the dependence of the RF magnetic permeability tensor on the static magnetic field.

The Static Field

As a general statement, the internal static field $\mathbf{H}_0$ is the solution of the static equations inside the ferrite with the appropriate constitutive relations and boundary conditions. In the simplest case of a saturated isotropic ferrite, the *static* constitutive relations reduces to $\mathbf{B}_0 = \mathbf{H}_0 + 4\pi\mathbf{M}_s$ (remember that we will use c.g.s. units in this part of the analysis), where $\mathbf{M}_s$ is the magnetization of the ferrite at saturation which, in isotropic ferrites, will be parallel to both $\mathbf{B}_0$ and $\mathbf{H}_0$. In many cases, the ferrite is placed in a known *external* static and uniform field $\mathbf{H}_{ext}$ provided by a magnet. In this case the internal static field is the sum of the external field and a demagnetization field $\mathbf{H}_d$ created by the ferrite internal magnetization. For ellipsoidal ferrite samples, rods and plates this problem is a classical one and is solved analytically, expressing the demagnetization field as the dot product of the saturation magnetization by a known demagnetization tensor, which depends on the shape and the orientation of the ferrite sample. In particular, for ferrite plates and rods placed in an external magnetic field parallel to the rod axis or the plane of the plate, it is easy to show that $\mathbf{H}_0 = \mathbf{H}_{ext}$. For ferrite plates placed in an external magnetic field perpendicular to the plate, it is $H_0 = H_{ext} - 4\pi M_s$.

The RF Magnetic Permeability Tensor

In this section we will state the more usual RF constitutive relationships for magnetized ferrites. First we will consider the simplest case of an intrinsically isotropic ferrite. In this context *intrinsic isotropy* means that the internal static magnetic field is the only source of anisotropy (induced anisotropy). Thus, after magnetization, the ferrite becomes an uniaxial medium with an RF permeability tensor given by (the z—axis is chosen as the direction of internal magnetization):

$$[\mu] = \mu_0 \begin{pmatrix} [\mu]_t & 0 \\ 0 & \mu_z \end{pmatrix} = \mu_0 \begin{pmatrix} \mu & jk & 0 \\ -jk & \mu & 0 \\ 0 & 0 & \mu_z \end{pmatrix} \tag{1}$$

where μ and κ are in general complex quantities that depend on the internal magnetic field and ferrite magnetization. This particular form of both $[\mu]$ and $[\mu]_t$ ensures the invariance of the permeability tensor after rotations around the z axis, which is the unique symmetry requirement. Notice that, if $\kappa \neq 0$, the tensor is not symmetric. Thus, a magnetized ferrite is a nonreciprocal medium. For lossless ferrites the permeability tensor must be hermitian and therefore μ, κ and μ_z are real numbers.

The theory leading to appropriate expressions for μ, κ and μ_z, valid for intrinsically isotropic saturated ferrites was first developed by Polder in 1949 from the analysis of the precession of molecular magnetic dipoles in the static internal field. The derivation of such expressions may be found in many textbooks [2] - [7]. The final expressions are:

$$\mu = 1 + \frac{\omega_M \omega_H}{\omega_H^2 - \omega^2} \tag{2}$$

$$\kappa = \frac{\omega\omega_M}{\omega_H^2 - \omega^2} \tag{3}$$

and $\mu_z = 1$, where ω_H is the resonance frequency given by

$$\omega_H = \gamma H_0 \tag{4}$$

and

$$\omega_M = 4\pi\gamma M_s \tag{5}$$

(γ is the gyromagnetic ratio $\gamma = \frac{ge}{2m_e c}$,, where g is the Lande factor, $-e$ and m_e the electron charge and mass respectively and c the speed of light; in most ferrites the magnetization is due to the electron spin alone, therefore $g = 2$ and $\gamma = 1.76 \times 10^7\ rad/\sec O_e$).

There is also a wide class of useful microwave devices which uses ferrites at the rema-nent magnetization. These are named *latch* ferrite devices and use ferrite materials with a square hysteresis loop, so that the remanence magnetization M_r is very close to the saturation magnetization. At remanence $H_0 = 0$ and (2)and (3)simplify to $\mu = 1$ and $\kappa = -4\pi M_r/\omega$. Nevertheless, the use of these expressions is subject to some restrictions related to the magnetic losses that may appear at low values of $\mathbf{H}_0$ [6].

The lossless ferrite is an approximation. Actually, there are many mechanism of losses in ferrites. Some of them, such as the ohmic conductivity, can be introduced in the RF constitutive relationships adding an imaginary part to the dielectric permittivity. Moreover, there are magnetic losses coming from the damping of the magnetic oscillations described by the Polder tensor. The presence of the resonance frequency ω_0 in that tensor clearly suggest the presence of losses in actual ferrites, with a maximum at this frequency. Magnetic losses may be included in the Polder tensor after the transformation [10]:

$$\omega_H \to \omega_H + j\alpha\omega \tag{6}$$

where α is a new parameter accounting for losses. The α parameter is often substituted by the *resonance linewidth* ΔH, i.e. the width of the resonance curves for the real part of κ and $\mu - 1$ plotted against H_0. The resonance linewidth is related to α by $\Delta H = 2\alpha\omega_H/\gamma$. The magnitude of magnetic losses varies widely in ferrites used for microwave applications. The resonance linewidth ranges from about 0.1 *Oe* for single YIG crystals to several hundreds for polycristalline ferrite materials. In the first case we can obtain meaningful results neglecting magnetic losses, but this approximation may lead to significant misleading in other cases.

Until now we have considered only ferrites with intrinsic isotropy. This assumption is not realistic in all cases, because ferrites are crystalline materials with complex internal structure. Magnetocrystalline intrinsic anisotropy usually induces in the crystal easy and hard directions of magnetization. The modifications of the Polder tensor induced by the magnetocrystalline anisotropy are complex and will not be analyzed here. The reader interested in this topic is referred to the textbooks that develop such expressions (e.g. [8]). The most important magnetocrystalline effect occurs in uniaxial hexagonal ferrites. In these crystals, the resonance is pushed to a extremely high frequency, even when the applied static field is small. Thus, hexagonal ferrites found most of its practical applications at millimeter wave frequencies.

The Radiofrequency Field

The RF field in a ferrite, characterized by the tensor permeability (1), is obtained by solving the Maxwell equations with the appropriate boundary conditions. Assuming a time-harmonic dependence of the kind exp $j\omega t$, as well as the vanishing of current sources, these equations will read:

$$\nabla \times \mathbf{E} = -j\omega[\mu] \cdot \mathbf{H} \tag{7}$$

$$\nabla \times \mathbf{H} = j\omega\epsilon\mathbf{E} \tag{8}$$

where the dielectric permittivity ϵ is, in general, a complex quantity, in order to incorporate dielectric and ohmic losses.

Uniform Plane Waves with Longitudinal Magnetization. Although the aim of this text is to analyze propagation along waveguides, the analysis of the propagation in an unbounded ferrite medium will provide an useful introduction to some relevant aspects of the propagation in ferrite waveguides. We will first suppose an uniform plane wave with a space-time dependence of the kind exp $j(-kz + \omega t)$ (this factor will be suppressed in the following) and an internal static magnetic field directed along the z—axis $\mathbf{H}_0 = H_0\boldsymbol{a}_z$. The analysis of (7)and (8)with these restrictions leads to two TEM wave solutions with right handed circular polarization (RCP) and left handed circular polarization (LCP) referred to the static field $\mathbf{H}_0$ orientation. These two waves have different phase constants given by:

$$k^{\pm} = \omega\sqrt{\epsilon\mu_{eff}} \tag{9}$$

where the + sign stands for the RCP polarization and the − sign for the LCP polarization and μ_{eff} is an *effective magnetic permeability* given by the two eigenvalues of $[\mu]_t$:

$$\mu_{eff}^{\pm} = \mu_0(\mu \pm \kappa) \tag{10}$$

The polarization handedness is defined with regard to the internal static magnetic field orientation, regardless of the direction of propagation. Therefore, if one of these waves is fully reflected (by a perfect conducting plate perpendicular to propagation, for instance), the handedness of the circular polarization, as well as the value of the propagation constant, will remain unchanged, giving rise to an stationary wave.

The values of $k^{\pm}$ for an isotropic lossless ferrite with RCP and LCP polarization are shown in Fig. 1 for a ferrite magnetized under the usual technological condition of $H_0 < 4\pi M_0$. A forbidden frequency range for RCP waves, in which k^+ becomes imaginary, is given by:

$$\omega_H < \omega < \omega_H + \omega_M \tag{11}$$

If magnetic losses are considered, the transformation (6)must be introduced in the expressions for $\mu_{eff}^{\pm}$. This leads to two complex propagation constants, the propagation constant of the RCP wave showing a typical resonant behavior with high resonance losses (see Fig. 1).

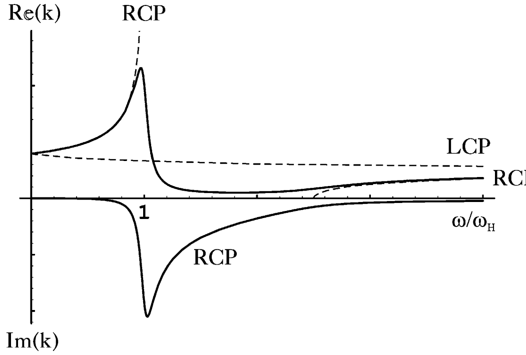


Figure 1. Normalized complex propagation constant for the RCP wave (10)in a lossy ferrite with $\omega_M = 1.5\omega_H$ and $\gamma\Delta H = 0.1\omega_H$ (solid lines). The normalized phase constants for the two RCP and LCP waves in a lossless ferrite with $\omega_M = 1.5\omega_H$ and $\gamma\Delta H = 0$ are also shown (dashed lines).

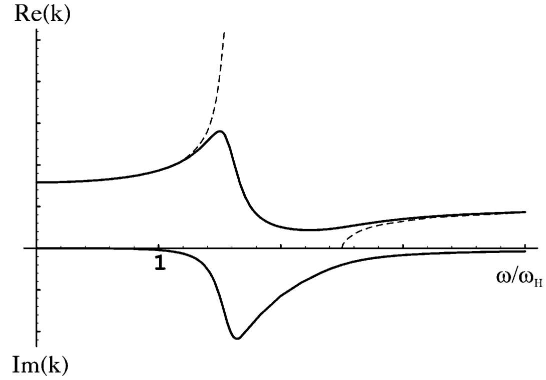


Figure 2. Normalized phase and complex propagation constants for the extraordinary waves in the infinite lossless (dashed lines) and lossy (solid lines) ferrites of Fig. 1 with transverse magnetization.

The most relevant effect related to plane wave propagation in a longitudinally magnetized ferrite is the non-reciprocal Faraday rotation of the plane of polarization of a linearly polarized wave. A linearly polarized wave is not a solution of (7)and (8), but it can be obtained by adding two contra-rotating RCP and LCP waves of equal amplitude. Since the phase constants of these two waves are not equal, the result is a rotation of the plane of polarization of the linearly polarized wave. The rotation angle after the wave has advanced a length Δz is given by:

$$\theta = \frac{1}{2}(k^- - k^+)\Delta z \tag{12}$$

When a linearly polarized wave is reflected backward, the hand of rotation of the polarization plane remains unchanged. Thus, the planes of polarization of the incident and the reflected waves will be different at a given distance from the plane of reflection. Therefore, the Faraday rotation in ferrites is non-reciprocal. If losses are considered, the RCP and the LCP waves have different attenuation constants. This leads to an unequal change in the amplitudes of the RCP and LCP waves, which causes Faraday elipticity of the original linearly polarized wave. Detailed treatments of Faraday rotation and elipticity may be found in the literature cited in the introduction.

Transverse Magnetization. Let us suppose now an uniform plane wave with a space-time dependence of the kind $\exp\ j(-kx + \omega t)$ (this factor will be suppressed in the following) and an internal static magnetic field directed along the z—axis $\mathbf{H}_0 = H_0\boldsymbol{a}_z$. The solution to (7)and (8)with these restrictions leads two independent uniform plane waves withe the **E** field linearly polarized. One of them is a TEM wave with the magnetic field parallel to $\mathbf{H}_0$. Thus, there is no interaction between the RF field and the electronic spins, and the effective magnetic permeability is $\mu_{eff} = \mu_0\mu_z$. This solution is called the *ordinary* wave, with phase constant $k^2 = \omega\sqrt{\epsilon\mu_0\mu_z}$. There is also an extraordinary wave, whose propagation constant is still given by (9), but with μ_{eff} given by:

$$\mu_{eff} = \mu_0\frac{\mu^2 - \kappa^2}{\mu} \tag{13}$$

The extraordinary wave is a TE wave with the electric field polarized parallel to $\mathbf{H}_0$ and the magnetic field elliptically polarized in the plane perpendicular to $\mathbf{H}_0$. The values of the propagation constants of the extraordinary waves for the lossless and the lossy ferrites of Fig. 1 are shown in Fig. 2. For the lossless extraordinary wave there is a frequency forbidden range for which k becomes imaginary, defined by:

$$\sqrt{\omega_H(\omega_H + \omega_M)} < \omega < \omega_H + \omega_M \tag{14}$$

The presence of an ordinary wave and an extraordinary wave with orthogonal polarization, recalls the birrefringence of uniaxial crystals. This birrefringence can be used in the design of microwave devices such as half and quarter wave plates, polarizers, etc.

Magnetization at Any Angle. In this case, the phase constant can still be written as in (9), with an effective magnetic permeability that depends on the angle θ_k between the static magnetization and the wave phase velocity. The final expression for this effective permeability is [10]:

$$\mu_{eff}(\theta_k) = \mu_0\frac{2 + (\frac{\mu_t}{\mu_z} - 1)\sin^2\theta_k \pm \sqrt{(\frac{\mu_t}{\mu_z} - 1)^2\sin^4\theta_k + 4\frac{\kappa^2}{\mu^2}\cos^2\theta_k}}{2(\frac{\sin^2\theta_k}{\mu_z} + \frac{\cos^2\theta_k}{\mu})} \tag{15}$$

with $\mu_t = (\mu^2 - \kappa^2)/\mu$. For a fixed frequency, a plot of k, θ_k in polar coordinates, with $k = \omega\sqrt{\varepsilon\mu_{eff}}$, gives the phase constants corresponding to the two solutions of (15). These curves can are the isofrequency curves of the dispersion equation $\omega = \omega(k, \theta_k)$ in the k, θ_k plane. The group velocity $\mathbf{v}_g = \nabla_k\omega$ is perpendicular to these curves at each point. Therefore, for arbitrary θ_k the direction of the optical ray is not parallel to the direction of propagation of the wave fronts.

Nonreciprocity. One of the basic theorems of electromagnetism is the Lorentz Reciprocity Theorem. It applies to any linear and causal media whose constitutive relationships can be described by symmetrical frequency dependent dielectric permittivity and/or magnetic permeability tensors. As it was already mentioned, since the tensor magnetic susceptibility (1)is not symmetric, this is not the case for ferrites. In fact, many of the practical applications of ferrites in microwave technology, such as circulators or isolators, arise from this non-reciprocal behavior. However, there is still possible to re-formulate the reciprocity theorem in a form that is applicable to ferrite media. Following Harrington [16] and McIsaac [17], we will start from Onsager symmetry relations, which states that any tensor macroscopic susceptibility of a causal and linear medium must be equal to its transpose after reversal in time of all the physical relevant quantities. For an externally magnetized ferrite, taking into account that the static bias field changes of sign after reversal in time, we conclude that the tensor magnetic susceptibility of a ferrite (1)must equal to its transpose after a change of sign of the static biasing field. From this conclusion, we can directly state the reciprocity theorem for ferrite media:

$$\iint (\mathbf{E}\times\mathbf{H}' - \mathbf{E}'\times\mathbf{H})\cdot d\mathbf{S} = \iiint (\mathbf{E}'\cdot\mathbf{J} - \mathbf{E}\cdot\mathbf{J}')dV \quad (16)$$

where the physical quantities must be reinterpreted as follows: $\mathbf{E}$, $\mathbf{H}$ and $\mathbf{E}'$, $\mathbf{H}'$ are two independent electromagnetic field configurations produced by source current densities $\mathbf{J}$ and $\mathbf{J}'$ respectively, at frequency ω in the same medium containing ferrites, except that the medium in which the prime quantities are defined has reversed static magnetization: $\mathbf{H}_0' = -\mathbf{H}_0$. The surface integrals on the left hand side of (16)are taken over any surface containing the source current densities included in the volume integral on the right side. This generalized reciprocity theorem is useful for the analysis of mode orthogonality in ferrite loaded waveguides, as well as in the analysis of ferrite loaded waveguide junctions.

MICROWAVE PROPAGATION IN FERRITE LOADED WAVEGUIDES

In the preceding section microwave propagation in unbounded ferrite media has been analyzed. Many of the studied effects, such as Faraday rotation and nonreciprocity, also appear when the RF field propagates along waveguides. Moreover, the microwave propagation along ferrite loaded waveguides presents new interesting and useful effects, such as unidirectionality, field displacement, propagation of complex and backward modes, slow magnetostatic waves, and many others.

In the following, we will choose the z—axis as the waveguide axis, and a space time dependence of the kind exp $j(\omega t - kz)$ will be supposed. The mode phase constant k will be, in general, a complex number $k = \beta - j\alpha$. Both β and α will be chosen real without loss of generality, and the factor exp $j(\omega t - kz)$ will be suppressed in the following.

Unidirectional and Bidirectional Modes

A mode with phase constant k which has not a symmetrical pair with the opposite phase constant $-k$ is called *unidirectional*. All lossless and reciprocal waveguides are bidirectional. This is not the case for ferrite loaded waveguides, because magnetized ferrites are nonreciprocal media which are not invariant after time reversal. The presence of unidirectional modes of propagation in ferrite loaded waveguides is useful in many microwave devices, such as isolators and nonreciprocal phase shifters. However, ferrite loaded waveguides which remain invariant after some symmetry transformations are bidirectional, i.e. unidirectional modes can not propagate along these waveguides. McIsaac [18] and, more recently, Dmitriyev [19] have investigated these symmetries. Mc.Isaac [18] concludes that bidirectionality is ensured if the waveguide remains the same after one or more of the following transformations:

- Reflection in a plane perpendicular to z axis.
- Rotation by 180° about an axis perpendicular to the z axis.
- Inversion at any point

In performing these transformations, the pseudo-vectorial nature of the static bias field $\mathbf{H}_0$ has to be taken into account (i.e. $\mathbf{H}_0$ remains the same after spatial inversion and after reflection in a a perpendicular plane, but changes of sign after reflection in a parallel plane). In particular, any ferrite loaded waveguide with longitudinal magnetization must be bidirectional, because this waveguide remains unchanged after reflection in a plane perpendicular to the z axis

Bidirectionality does not imply that all the characteristics of the modes remain unchanged when the direction of propagation is reversed. For instance, the energy distribution and/or the polarization of a pair of bidirectional modes having the same but opposite phase constant may be different. Moreover, although all modes in bidirectional ferrite loaded waveguides must be bidirectional, not all the modes in non bidirectional ferrite loaded waveguides are unidirectional: Some of them, having the appropriate polarization, may be bidirectional.

Complex and Backward Modes

Complex modes in inhomogeneously filled lossless waveguides were first reported by Tai in 1960 and by Carricoats in 1965 [20]. Complex modes in lossless reciprocal waveguides are characterized by a complex propagation constant $k = \pm\beta \pm j\alpha$ and appear in groups of four solutions, for which all the possible combination of signs are allowed. However, in non reciprocal waveguides, unidirectional complex modes with $k = \beta \pm j\alpha$ may appear. For a single complex mode, power flows in opposite directions along the different media filling the waveguide, giving a zero net power flux. Therefore, complex modes in lossless waveguides are reactive modes. Complex modes have proven to be a very important part of the spectra of ferrite loaded waveguides [20] (in fact they were first reported in ferrite loaded waveguides by Tai). In particular, all the unidirectional and reactive modes in ferrite loaded waveguides must be complex [21, 22].

Complex modes are closely related to *backward modes* (i.e. modes with negative group velocity). In fact, a pair of complex modes in lossless waveguides usually changes to a pair of propagating *forward* and *backward* modes when frequency varies [20]. Backward modes in the spectra of ferrite loaded waveguides, mainly in the *magnetostatic wave* region, have been widely analyzed (see, for instance, [8]).

Mode Orthogonality

Mode orthogonality in ferrite loaded waveguides was analyzed in [23]. Applying the generalized reciprocity theorem (16)to two modes $b\ fe_m, \mathbf{h}_m(x, y) \exp\ j(\omega t - k_m z)$ and $b\ fe'_n, \mathbf{h}'_n(x, y) \exp\ j(\omega t - k'_n z)$ of the actual waveguide and the *complementary* waveguide (the complementary waveguide is defined as the original one with the static magnetic field reversed), the following relation is obtained:

$$(k'_n + k_m) \iint (\mathbf{e}'_n \times \mathbf{h}_m - \mathbf{e}_m \times \mathbf{h}'_n) \cdot \mathbf{u}_z\, dxdy = 0 \qquad (17)$$

where the integral must be taken over the cross section of the waveguide. This equation can be considered as a general orthogonality relation since the integral must be zero unless $k_m = k'_n$. This relation simplify for most practical situations, in which the static magnetic field is either parallel or perpendicular to the waveguide axis. The explicit orthogonality relationships for these particular but important cases can be found in [23].

Field Displacement Effects

It was pointed out above that electromagnetic wave propagation in unbounded ferrites at an oblique angle with respect to the magnetizing field usually implies that power flux and phase velocity are not parallel. However, in a nonradiating waveguide, both power flux and wave propagation are forced to be parallel to the waveguide axis. Thus, this effect can not be present in non-radiating waveguides. Instead, this tendency of energy to flow in a direction different from wave propagation may cause strongly unsymmetrical accumulations of electromagnetic energy across the waveguide section. This effect is usually nonreciprocal and can be used in the design of microwave isolators and phase shifters.

The Magnetostatic Approximation. Near the resonances $\mu_{eff} \to \infty$ and the effect of the Maxwell displacement current may be neglected with regard to Faraday induction effects. This leads to the magnetostatic approximation. Taking into account that $\nabla \cdot \mathbf{B} = 0$, a magnetostatic potential $\mathbf{H} = -\nabla\psi$ is defined, which must satisfy:

$$\nabla \cdot ([\mu] \cdot \nabla\psi) = 0 \qquad (18)$$

(in ferrite loaded waveguides, the nabla operator is replaced by $\nabla \to \nabla_t - jk\mathbf{u}_z$. The solutions to (18), with the appropriate boundary conditions, are the magnetostatic modes of the waveguide. Magnetostatic surface and volume wave propagation in layered ferrite loaded structures is extensively analyzed in [8]. Magnetostatic waves can be also excited in microstrip and slot lines [24]. The main applications of magnetostatic waves in microwave technology arises from its small wavelength. This result in broad applications in miniature controllable devices, such as delay lines, filters, power limiters and signal to noise enhancers [25].

Basic Properties of Ferrite Loaded Waveguide Junctions

A waveguide junction is characterized by its scattering matrix $S_{i,j}$. It is a well known fact that, if the materials filling the junction are reciprocal, the scattering matrix must be symmetrical. If the junction is nonreciprocal, this statement must be modified as a consequence of the reformulation of the reciprocity theorem (16). This modification leads to the following relations between the scattering matrix elements of a ferrite loaded junction and its complementary (i.e., the junction with the biasing static magnetic field reversed):

$$S_{i,j}(\omega, \mathbf{H}_0) = S_{j,i}(\omega, -\mathbf{H}_0) \qquad (19)$$

If the junction is also lossless, the scattering matrix must be unitary ($S_{i,j} S^{*}_{k,j} = \delta_{i,k}$; where the rule of summation over all the repeated subindex has been used). Other symmetries of the scattering matrix may be deduced from the spatial symmetries of the junction (including the bias field) [19].

The use of the scattering matrix symmetry properties is useful in the design of many microwave devices, such as isolators, phase shifters and circulators. The *Y circulator* is perhaps the most useful and known nonreciprocal junction. An Y circulator is a symmetrical three port junction with some specific properties. A symmetrical three port junction must have $S_{1,1} = S_{2,2} = S_{3,3}$, $S_{1,2} = S_{2,3} = S_{3,1}$ and $S_{2,1} = S_{1,3} = S_{3,2}$. These relations are fulfilled by any junction having a rotation symmetry axis of third order and magnetized along this axis. The circuit theory of three and N port circulators may be found in [6] and other textbooks. It can be shown that if a lossless, nonreciprocal and symmetrical three port waveguide junction is matched (i.e. $S_{1,1} = 0$), it is also an ideal Y circulator (i.e. $S_{1,2} = 1$ or 0, and $S_{1,3} = 0$ or 1). If the magnetization of a nonreciprocal three port Y circulator is reversed, the direction of circulation is also reversed, as a consequence of (19).

FERRITE LOADED WAVEGUIDES FOR PRACTICAL APPLICATIONS

In this section we will describe the most widely used ferrite loaded waveguides. There are many classical textbooks and papers, eg. [5], [2], [3], [6], [7], and more recently [9], [26]... that describe these waveguides, as well as the most useful microwave devices that may be designed using them. The reader may use these and other texts for broadening the information contained in this section.

Circular Waveguides with Longitudinal Magnetization

It is a well known fact of the theory of hollow waveguides that the fundamental mode of the empty circular waveguide is the $TE_{1,1}$ mode, which is a double degenerate mode with two perpendicular polarizations in the waveguide cross section. This mode has a field distribution which is almost TEM in the vicinity of the waveguide axis. Thus, if a ferrite rod with longitudinal magnetization is placed at the center of the waveguide (see Fig. 3), the two orthogonal and degenerate $TE_{1,1}$ fundamental modes will interact as a consequence of the Faraday rotation effect, giving

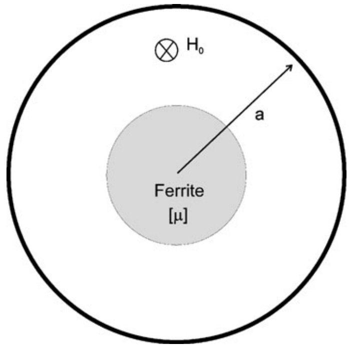


Figure 3. Cylindrical waveguide with an inner centered ferrite rod of longitudinal magnetization.

rise to two nondegenerate cicularly polarized RCP and LCP modes (with the hand of polarization defined with respect to the static bias field orientation). These modes can be approximated by the same RCP and LCP modes of a circular waveguide with an inner isotropic rod with an scalar magnetic susceptibility given by (10). In the same way, if a linearly polarized wave with linear polarization enters the ferrite loaded waveguide, this wave will experience a Faraday rotation by an angle approximately given by (12), where k^+ and k^- are now the phase constants of the RCP and LCP ferrite loaded waveguide modes.

The phase constants, as well as the mode fields of the two nondegenerate RCP and LCP ferrite loaded waveguide modes were obtained analytically by Waldron in 1958. Analytical solutions, not only for the cylindrical waveguide with a ferrite rod, but also for many other related structures, such as cylindrical waveguides loaded with ferrite and dielectric tubes, may be also found in [4]. Modes in this kind of ferrite loaded waveguides are not *TE* nor *TM*, but becomes *TE* ad *TM* at cutoff [4], therefore modes are named *HE* and *EH* depending on whether the magnetic H_z or the electric E_z field dominates. At cutoff, *HE* modes become *TE* and *EH* modes become *TM*.

RCP modes in ferrite filled circular waveguides present an interesting behavior. For such modes, the effective transverse permeability is approximately given by (10), becoming negative in some frequency band (see Fig. 1). However, the longitudinal permeability is $\mu_0 > 0$. This combination of negative/positive transverse/longitudinal permeability gives rise to an anti-cutoff behavior [27]: the RCP mode is evanescent when the empty waveguide is above cutoff, and becomes propagative when the waveguide radius decreases, so that the empty waveguide is at cutoff. This anti-cutoff behavior appears for the RCP mode in the region of negative μ^+. Outside this frequency band the mode presents a regular behavior, becoming evanescent when the waveguide radius decreases.

Ferrite loaded circular waveguides with longitudinal magnetization are extensively used in Faraday rotation devices, based on the aforementioned rotation of the polarization plane of a linearly polarized wave. The most known Faraday rotation device is the four port circulator, described in many textbooks. Faraday rotation may be also used in the design of magnetically tuned variable attenuators, isolators and phase shifters (see for instance [9] and references therein). In the first years of the microwave ferrite technology, much effort was devoted to develop Faraday rotation circulators and other microwave devices with cylindrical geometry. In the following years, however, the *Y*-junction circulators, as well as phase shifters and attenuators in rectangular and/or planar technology were found to be smaller, simpler and more appropriate for most applications, and the research effort turns on these devices.

E-plane Transversely Magnetized Ferrite Loaded Rectangular Waveguides

Fig. 4.a shows the variation of the magnetic field components H_x and H_z of the fundamental $TE_{1,0}$ mode in a hollow rectangular waveguide. The magnetic field is circularly polarized around the y axis were $|H_z| = |H_x|$. This condition occurs at two symmetrical positions, at a distance d of the rectangular side walls, given by:

$$d = \frac{a}{\pi}\cot^{-1}(\sqrt{\frac{4a^2}{\lambda_0^2}} - 1) \tag{20}$$

If an E-plane ferrite slab biased with a static magnetic field directed along the y axis is placed at a distance d of one of the lateral side walls (see Fig. 4.b), the wave propagating in the positive (negative) direction along the z-axis is right (left) handed polarized with respect to the static bias field. Thus, we can expect that the forward (backward) wave will *see* the effective magnetic permeability of the slab μ^+_{eff} (μ^-_{eff}), given by (10). Therefore, wave propagation will be unidirectional, with different phase constants for the opposite directions of propagation. Moreover, since the dependence of μ^+_{eff} with $\mathbf{H}_0$ is much stronger than that of μ^-_{eff}, the forward wave will be much more affected by variations in the intensity of $\mathbf{H}_0$ than the backward one. If the bias field is chosen so that μ^+_{eff} is positive the waveguide can be used as a nonreciprocal phase shifter. If the bias field is chosen at the resonance condition ($\omega_H = \omega$), the forward wave will see a resonant magnetic permeability and will experience strong attenuation due to the resonance losses. Then, the waveguide may be used as a resonance isolator. If the bias field is chosen at the antirresonance condition $\omega_H = \omega - \omega_M$, the forward wave will see a perfect diamagnet with $\mu^+_{eff} = 0$, which imposes perfect diamagnetic boundary conditions at both slab sides and, therefore, zero tangential electric RF field at these boundaries. If an absorber is located at the inner boundary of the slab, it is expected that the forward wave will not be attenuated whereas the backward wave will be strongly attenuated. This configuration can be used as a field displacement isolator.

The wave propagation characteristics along this waveguide can be found analytically. The first published results on this subject are due to of Kales (1953). Gardiol [28] gave a general method for computing the propagation characteristics of rectangular waveguides filled with an arbitrary number of anisotropic slabs - including ferrite slabs - making use of the transverse transmission matrix method. Referred to the geometry of Fig. 4.b, the transverse transmission matrix of the i-region [$\mathbf{T}_i$] is defined as the matrix relating the tangential fields, E_y and H_z, at both sides of

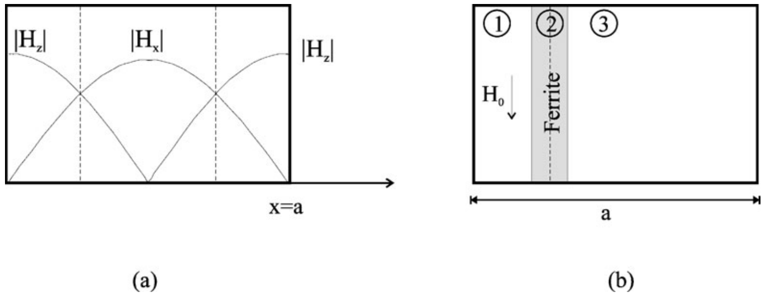


Figure 4. (a) Plot of the intensities of the magnetic field components $|H_x|$ and $|H_z|$ of the fundamental $TE_{1,0}$ mode in a rectangular hollow waveguide (frequency 9 GHz, a=23mm.), dashed lines: planes of circular polarization. (b) Rectangular waveguide loaded with a transversely magnetized ferrite slab at a plane of circular polarization of the RF magnetic field

the *i-th* region of the waveguide. In a notation that becomes apparent, we can write:

$$\begin{aligned}\begin{pmatrix} E_y \\ H_z \end{pmatrix}_{x=a} &= [T_3]\cdot[T_2]\cdot[T_1]\cdot\begin{pmatrix} E_y \\ H_z \end{pmatrix}_{x=0} \\ &= \begin{pmatrix} t_{1,1} & t_{1,2} \\ t_{2,1} & t_{2,2} \end{pmatrix}\cdot\begin{pmatrix} E_y \\ H_z \end{pmatrix}_{x=0} \end{aligned} \tag{21}$$

Since the lateral side walls are assumed to be perfect conducting walls, the tangential electric field must vanish at these boundaries. This imply that $t_{1,2} = 0$, which can be considered an implicit equation for the phase constant. This method can be applied to rectangular waveguides with any number of E-plane transversely magnetized ferrite slabs and/or lossy dielectric slabs, thus providing a general method for the analysis of ferrite isolators and phase shifters in rectangular waveguide technology.

The nonreciprocal isolation and phase variation effects of the E-plane ferrite loaded waveguide of Fig. 4.b may be increased by placing a symmetrical ferrite slab with reverse magnetization at the remaining plane of circular polarization of the $TE_{1,0}$ mode. Since at this plane the circular polarization of the $TE_{1,0}$ wave has opposite handedness, the effect of the new ferrite slab adds to the effect of the former one. A variation of this two-slabs ferrite loaded waveguide is the latch ferrite toroid in rectangular waveguide (Fig. 5.a) proposed by Treuhaft (1958) for phase sifting applications. The main advantage of this configuration is that the permanent magnet is substituted by a ferrite toroid magnetized at remanence by an electric current pulse, driven by a single wire at the center of the waveguide (this wire is perpendicular to the RF electric field and has a negligible effect on microwave propagation). This structure is also suitable for fast switching between the two opposite nonreciprocal states of the waveguide. In the analysis of this structure, the upper and lower branches of the ferrite toroid, which do not have substantial effect in phase change, may be neglected leading thus to the simpler structure of Fig. 5.b. Gardiol [29] give expressions that transform the geometry of Fig. 5.a in the geometry of Fig. 5.b with a gain in accuracy. An alternative for reducing the unwanted effects of the upper and lower branches of the ferrite toroid is to place this into a rectangular grooved waveguide, as proposed in [30]. If nonreciprocity is not desired, a reciprocal phase shifter may be still obtained magnetizing both slabs of Fig. 5.b with parallel and equal static magnetic field. This structure is symmetric after inversion at a point in the waveguide axis and, therefore, it is bidirectional.

Regarding the anti-cutoff behavior previously reported for ferrite-filled waveguides with longitudinal magnetization [27], a similar behavior has been recently reported in ferrite filled rectangular waveguides with transverse magnetization [31]. Since both configurations show backward-wave characterististics, they seem to be good candidates for left-handed media simulation in a waveguide environment [32].

Other Useful Cylindrical and Rectangular Ferrite Loaded Waveguides

Although the circular and the rectangular geometries seems to be the natural geometries for longitudinal and transverse magnetization, respectively, there are also some useful devices which use transversely magnetized circular waveguides and longitudinally magnetized rectangular waveguides. Similar effects to those reported above for rectangular waveguides are present in cylindrical waveguides loaded with latch ferrite tubes magnetized in the azimu-tal direction. The dual-mode ferrite phase shifters include latch and transversely magnetized circular waveguide sections [33]. A widely used ferrite loaded rectangular waveguide with longitudinal magnetization is the Reggia-Spencer phase shifter [34], which consists of a ferrite rod with longitudinal magnetization placed at the center of a rectangular waveguide. If the dimensions of the hollow waveguide only allows for the propagation of the first $TE_{1,0}$ mode, Faraday rotation can not take place. Instead, a strong variation of the wave phase constant with the applied static magnetic field occurs. Like all waveguides having longitudinal magnetization, the Reggia-Spencer phase shifter is bidirectional, therefore the phase shift is reciprocal.

FERRITE LOADED MICROSTRIPS, SLOT LINES AND FIN LINES

After the middle of the 1960s, when planar microwave integrated circuits became a viable technology, ferrite loaded microstrips and slot lines began to be investigated as an alternative to traditional ferrite loaded waveguides for the design of reciprocal and non reciprocal phase shifters [35], isolators [36] and other useful devices, which have been summarized in some classical review papers [26] and textbooks [9]. Later, when fin lines emerged as an useful alternative for planar technology in millimeter wave circuits, ferrite loaded fin lines [37], [38] also began to be investigated.

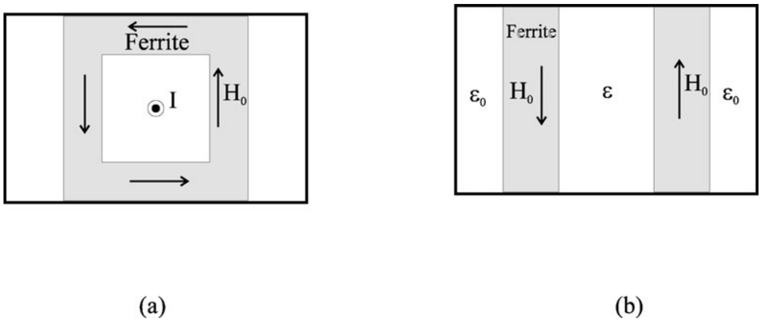


Figure 5. (a) Rectangular waveguide loaded with a latch ferrite toroid. (b) Rectangular waveguide loaded with two oppositely magnetized ferrite slabs at the planes of circular polarization of the r.f. magnetic field.

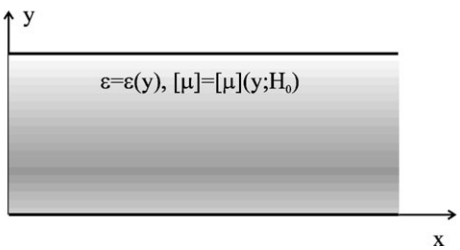


Figure 6. Parallel plate waveguide filled by a multilayer medium including one or more ferrite slabs ($\epsilon(y)$ and $[\mu](y)$ are piezewise constant functions of y)

The Ferrite Loaded Parallel Plate Waveguide

Before considering ferrite loaded standard planar transmission lines, it will be useful to analyze the much simpler ferrite loaded parallel plate waveguide of Fig. 6. It has been shown [26] that for magnetization parallel to propagation, these waveguides supports a quasi-TEM mode, and that the ferrite layers may be characterized by an effective permeability given by (13). For magnetization perpendicular to both, the direction of propagation and the plane of the waveguide, the ferrite layers may be again characterized by the scalar effective permeability (13), but the field is no longer TEM due to birrefringence effects. Finally, for magnetization perpendicular to propagation and parallel to the waveguide plane, there is almost no interaction between the static magnetic field and the RF field, and the ferrite layers are characterized by the scalar permeability μ_z. Near the forbidden frequency range, the analysis becomes more complicated, due to the apparition of magnetostatic modes. Magnetostatic modes in parallel plate ferrite loaded waveguides have been extensively analyzed in [8] and references therein.

Ferrite Loaded Microstrip Lines

Microstrip line (see Fig. 7.a) is the most used waveguide in planar technology. Although exact methods of analysis are now available, considerable insight on the physical behavior of ferrite loaded microstrip lines can be obtained from the well known parallel plate microstrip model (see Fig. 7.b). In this model, the microstrip line is substituted by a section of parallel plate waveguide between two magnetic walls. This section is slightly wider than the microstrip, in order to incorporate the effects of the fringing fields. The parallel-plate waveguide model for the microstrip line is valid provided there is not radiation leakage through surface waves excited at both sides of the microstrip.

Microstrip with Longitudinal Magnetization. Using the model of Fig. 7.b and the results reported above, the qualitative behavior of ferrite loaded microstrip lines can be deduced - at least qualitatively - from the analysis of a section of parallel plate waveguide loaded with one or more magnetic slabs with the effective magnetic permeability (13), bounded by two perfect magnetic walls. This model predicts a bidirectional quasi-TEM fundamental mode, with a phase constant which is a function of the biasing magnetic field. The same qualitative results are provided by more accurate quasi-TEM analysis of the actual microstrip line, using either the effective permeability (13)[26], or the tensor magnetic permeability (1) [39], [40]; or by a full wave analysis that will be discussed later in this article. The main application of microstrip lines with longitudinal magnetization is in phase shifting by meander lines, a design that minimize the size of the device [35]. The phase variation with the applied magnetic field may be increased if strongly coupled quarter wave meander line sections are used. These structures provides strong nonreciprocal phase shifting [35] were nonreciprocity is due to the coupling effects.

Transversely Magnetized Microstrip Lines. Along this section we will consider the two orthogonal magnetizations, perpendicular and parallel to the ground plane. For the second case, since the RF magnetic field is parallel to the static bias field, there is no interaction between the bias and the RF fields in the parallel plate waveguide model of Fig. 7.b. Therefore, a very small interaction, due only to fringing fields, is expected in the actual microstrip line. In fact, only a slightly nonreciprocal phase shift is observed in practical devices.

Of much more interest is the microstrip line with magnetization perpendicular to the ground plane. First, a single microstrip line over a grounded ferrite slab will be considered. Considering anew the parallel plate waveguide model of Fig. 7.b, Hines [36] showed that such waveguide supports a non-reciprocal (though bidirectional) quasi-TEM mode, with a strong field displacement towards one of the magnetic walls. For invititely wide strips, the phase constant of this mode is given by $\mathbf{k} = k_0(\sqrt{\epsilon_r \mu}\mathbf{a}_z + j(\kappa/\mu)\sqrt{\epsilon_r \mu}\mathbf{a}_x)$, and its imaginary part accounts for the field displacement. The sign of this imaginary part - and, therefore, the direction of the field displacement - changes when the direction of propagation, or the static magnetization, is reversed (interestingly, $\mathbf{k}\cdot\mathbf{k} = k_0^2\varepsilon_r(\mu^2-\kappa^2)/\mu$ i.e. $\mathbf{k}\cdot\mathbf{k}$ is the same as for a nonuniform plane wave in a slab of

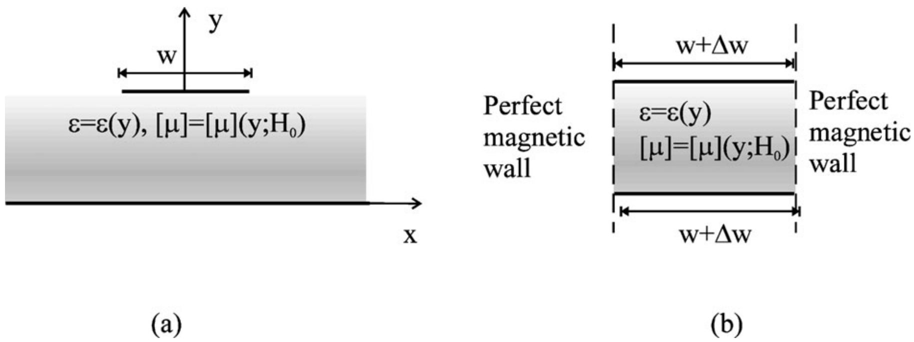


Figure 7. (a) Microstrip line on a multilayer ferrite loaded substrate (as in Fig. 6 $\epsilon(y)$ and $[\mu](y)$ are piezewise constant functions of *y)*. (b) Parallel plate waveguide model for the microstrip line of (a).

effective permeability μ_{eff} (13)). Although this analysis has been made for a specific configuration, similar qualitative results can be expected for multilayered ferrite loaded microstrip lines. Hines modes, also named edge modes because the RF field is mainly concentrated in the vicinity of a microstrip edge, are useful for the design of wideband edge mode isolators and nonreciprocal phase shifters [36], [42], [43]. For edge mode isolators, for instance, a small piece of absorbing material is placed near one of the microstrip edges, so as only one of the Hines modes - that with the appropriate direction of propagation - is attenuated. The analysis of this kind of structures is usually made by approximate models. More recently, the spectral domain integral method has been successfully applied to the analysis of edge mode isolators without approximations [44].

Ferrite Loaded Slot Lines and Fin Lines

Slot lines, Fig. 8.a, and coplanar waveguides are useful alternatives to microstrip lines in the design of microwave integrated circuits. In millimeter wave technology, fin lines in rectangular waveguides, Fig. 8.b, also are a good alternative for integration which prevents radiation losses. By adding ferrite layers to these waveguides, many of the described effects for ferrite loaded microstrips may be achieved. Since the RF magnetic field in slot-lines and fin-lines is mainly concentrated in the slot and directed perpendicular to the air interface, it is expected that the strongest effects for transversely magnetized slot-lines and fin-lines will occur for magnetization parallel to the substrate layers. Such kind of structures have been proposed for the design of field displacement isolators and phase shifters [26]. A millimeter wave field displacement fin line isolator was proposed in [37]. Transversely magnetized slot-lines and fin-lines for nonreciprocal phase shifting applications have been analyzed in [46], [47], [41] and [48]. Applications of fin-lines with longitudinal magnetization have been also investigated [38].

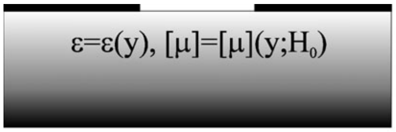
Methods of Analysis of Ferrite Loaded Quasi-Planar Layered Structures

With a few exceptions, quasi-TEM analysis usually provide accurate enough results for the analysis of conventional microstrip and coplanar or slot lines. However, this analysis is usually not suitable for these structures when they are ferrite loaded. In fact, quasi-TEM modes are by definition bidirectional and reciprocal, and therefore the quasi-TEM analysis can not take into account many of the most relevant physical effects in ferrite loaded transmission lines. Quasi-TEM analysis in its standard form is restricted to longitudinally magnetized lines [39], [40]. More recently, however, some attempts have been made in order to generalize this analysis to transversely magnetized structures [45]. Nevertheless, in general, planar and quasi-planar ferrite loaded transmission lines need of a full-wave analysis.

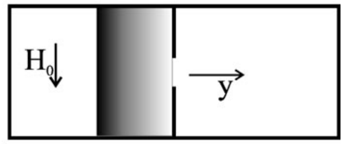
With regard to numerical techniques, spectral domain analysis (SDA) is by far the most widely used technique for the analysis of ferrite loaded strip and/or slot structures ([39], [40], [41], [44], [47], [48], [41] and [49]). Fundamentals of SDA may be found in many textbooks, as that of Misherkar-Shyakal [50]. SDA is specially well suited for the analysis of microstrips and/or slot- and fin-lines on planar single- or multilayer substrates, because of the translational symmetry of these substrates. Since the SDA applied to microstrip or microslot structures is adequately described in [50] and other textbooks, we only briefly describe here the main specific characteristics of the SDA when is applied to ferrite loaded microstrip and/or microslot waveguides. The main difficulty in the application of the SDA to ferrite loaded microstrip or microslot lines on infinite planar substrates is the determination of the spectral domain Green's function dyad $\overset{\circ}{\mathbf{G}}(k_x, k_z)$ which relates a surface current source $\mathbf{J}_s = \mathbf{J}_{s,0} \exp - jk_x x \exp - jk_z z$ in the plane of the structure, with the r.f. tangential electric field $\mathbf{E}_t = \mathbf{E}_{t,0} \exp - jk_x x \exp - jk_z z \, (\mathbf{E}_t = (E_x, E_z)^t)$ over the same or other parallel plane:

$$\mathbf{J}_0 = \overset{\circ}{\mathbf{G}}(k_x, k_z) \cdot \mathbf{E}_{t,0}, \tag{22}$$

General methods for the computation of the spectral Green's dyad in multilayered ferrite loaded substrates (in fact in general layered bianisotropic substrates) are reported in [51] and [52]. The SDA may be also applied to boxed strip- and fin-line structures. In this case the integral Fourier transform of the field and currents must be substituted by a Fourier series transform in an equivalent periodic structure. For magnetized ferrite layered media in rectangular metallic boxes this imposes an important restriction: Strictly speaking, the SDA can only be applied to substrates with static magnetization perpendicular to the lateral side walls. In any other case, due to the properties of the magnetic field after spatial reflection, it is not possible to find an equivalent periodic structure with translational symmetry suitable for the application of the series Fourier transform. Therefore, the application of the SDA to boxed



(a)



(b)

Figure 8. (a) Slot line line on a multilayer ferrite loaded substrate. (b) finline with a multilayer ferrite loaded substrate. As in Figs.6 and 87 $\epsilon(y)$ and $[\mu](y)$ are piezewise constant functions of y.

structures magnetized in any direction different from the aforementioned one, must be considered only as an approximation. The SDA, as described previously, only applies to structures with strips or fins of negligible thickness. Structures with non-negligible fin or strip thickness may be analyzed using a mode matching technique in the transverse direction, which also implies an SDA [53]. A similar method may be applied to boxed structures with asymmetrical rectangular piecewise boxes [54]. Finally, the SDA also applies to structures having fully or partly lossy strips or fins, provided that these lossy strips or fins can be described by a suitable surface impedance, defined over the strip or the fin region [44].

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