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Novel Microstrip-Line Directional Filters

Stanislaw Rosloniec and Tahar Habib

***Abstract*—This paper presents two new structures of four-port microwave directional filters incorporating microstrip-line resonators, uniplanar 180° phase-reversal units, and two or four identical p-i-n diodes. These filters provide substantially improved frequency characteristics compared to the conventional constructions. The unique feature of them is that they ensure high-level isolation independent of frequency between opposite ports. Due to this electrical property the proposed filters may be suitable for some pulse-radar applications.**

***Index Terms*—Microstrip-line directional filters, microwave filters, T/R modules.**

I. Introduction

Directional filters are indispensable components of transmit/receive (T/R) modules (see Fig. 1) used in various communication systems. The conventional nonadjustable structures of them are comprehensively described in the literature [1]–[3]. These filters, however, cannot be applied directly in systems with transmit and receive signals of the same or very close frequencies. Furthermore, they do not ensure perfect isolation independent of frequency between opposite ports marked as 2 and 3 in Fig. 1. Consequently, the receiver cannot be sufficiently protected from the transmitter over the required frequency range. In order to eliminate the above disadvantages two electronically switchable microstrip-line directional filters have been developed. The construction and electrical performance of these filters are the subject of this paper. The validity of the presented theoretical results has been confirmed experimentally for the frequency range of 0.5–1.5 GHz.

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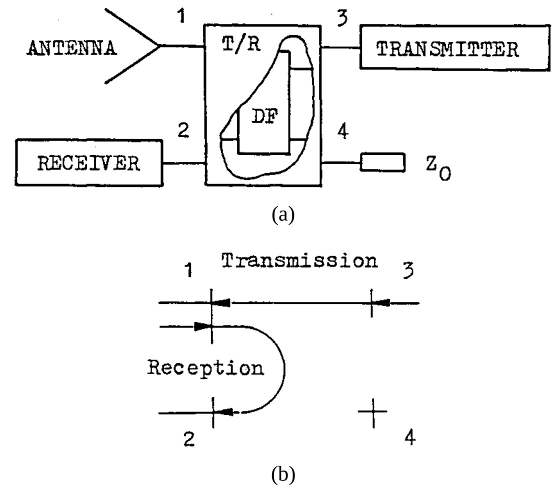


Fig. 1. A transmit/receive module. (a) Functional block diagram. (b) Signal flow chart

II. Microwave Directional Filters with Two and Four p-i-n Diodes

The microstrip line configurations of the filters being investigated are shown in Fig. 2. These filters are composed of two microstrip-line resonators, two or four p-i-n diodes, and a 180° phase-reversal unit which is placed at the center of one resonator. Near this center the standing wave forms a short-circuit node. Therefore, the equivalent inductances L of this unit [Fig. 3(a) and (b)], have to be especially small. This requirement explains why in the proposed filters relatively wide and short copper strips are used as electrical bonds. A constructional view of the proposed design, using the slotline hollow patches for broad-band open conditions, is illustrated in Fig. 3(c) and (d) [4], [5]. The second resonator including the 180° phase-reversal unit should be electrically uniform over the entire length, so that at any cross section its characteristic impedance will be the same. Thus, the finite ground plane of the microstrip line caused by the slotline hollow patches has to be taken into account in the design process.

The first filter (DF-2) presented in Fig. 2(a) is asymmetric with respect to the horizontal x–x' and vertical y–y' planes, and for this reason cannot be analyzed by using the even- and odd-mode excitation method [3], [6]. Consequently, the following numerical algorithm is proposed here for this purpose. The idea of this algorithm is similar to that used in [7]. For any pair of ports, k and l, the filter DF-2 may be treated as a reciprocal two-port network incorporating two partial two-ports P_{kl} and Q_{kl} connected in parallel, as shown in Fig. 4(a). For clarification of further considerations let us assume that the scattering parameters S_{11}, S_{14}, S_{41}, and S_{44} of the filter are evaluated. In this case, two-ports P_{14} and Q_{14} are similar to those shown in Fig. 4(b) and (c). The transfer matrices $(ABCD)_P$ and $(ABCD)_Q$ of these circuits can be easily calculated by multiplying the corresponding matrices $(ABCD)$ of the elementary cascade components [8]. When the resulting matrices $(ABCD)_P$ and $(ABCD)_Q$ are known, then we can evaluate the admittance matrices $[Y]_P$ and $[Y]_Q$ related to them. The total admittance matrix $[Y] = [Y]_P + [Y]_Q$ makes it possible to calculate the scattering parameters S being sought. For this purpose we can use the well-known matrix transformations $[Y] \rightarrow [S]$, [8], [9]. The

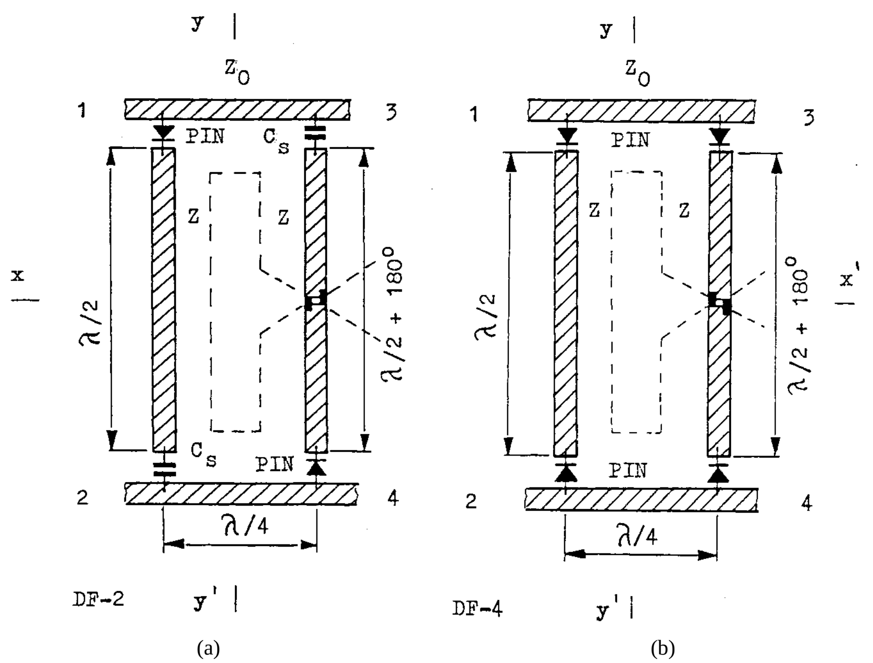


Fig. 2. Microstrip-line directional filters. (a) DF-2. (b) DF-4.

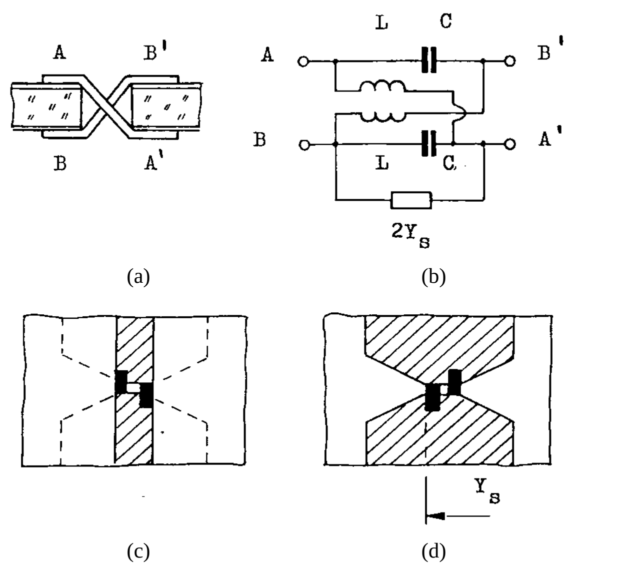


Fig. 3. A 180° phase-reversal unit. (a) Cross section. (b) Equivalent circuit. (c) Front view. (d) Rear view.

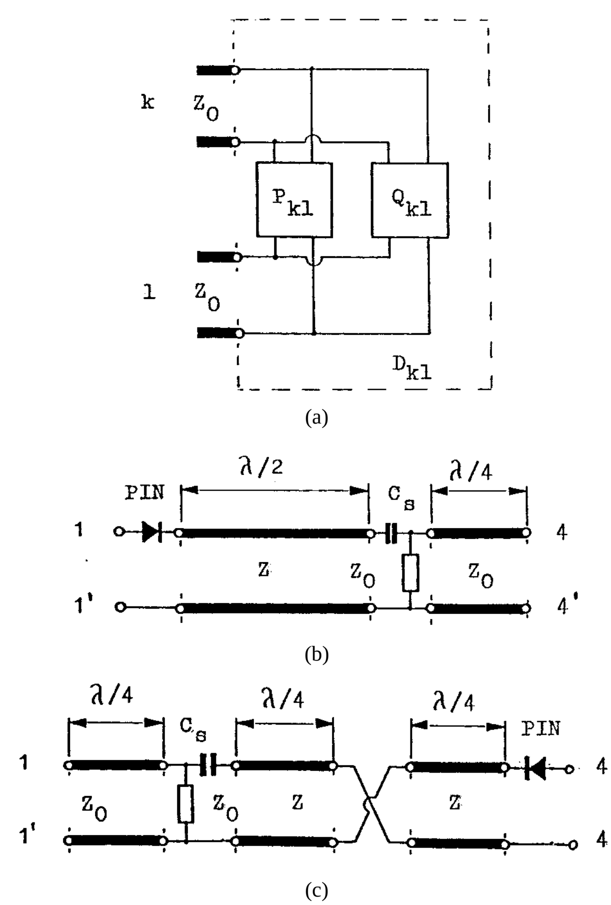


Fig. 4. Equivalent circuits for determining the S_{11}, S_{14}, S_{41}, and S_{44} scattering parameters. (a) General circuit. (b) Two-port P_{14}. (c) Two-port Q_{14}.

remaining scattering parameters of the filter may be evaluated in a similar manner. Then, we can calculate the voltage standing-wave ratio and insertion-loss characteristics expressed as

$$VSWR_k(f) = [1 + |S_{kk}(f)|]/[1 - |S_{kk}(f)|] \tag{1}$$

and

$$L_{kl}(f)[dB] = 20\log[1/|\mathrm{S_{kl}(f)}|] \tag{2}$$

where $k = 1, 2, 3, 4$ and $l = 1, 2, 3, 4$.

The filter DF-4 shown in Fig. 2(b) is symmetrical with respect to the horizontal plane $x = x'$ and can be analyzed by using the even- and odd-mode excitation method [6]. Of course, this one may be analyzed also in the manner described above. Such an approach has been used for calculating the frequency characteristics presented in Fig. 5. The calculations have been carried out for: $Z_0 = 50\ \Omega$, $Z = 50\ \Omega$, $C_s = 2.2$ nF and p-i-n diodes (chips) of type MA4P7000 forward biased (see Appendix). In this case, the reactance

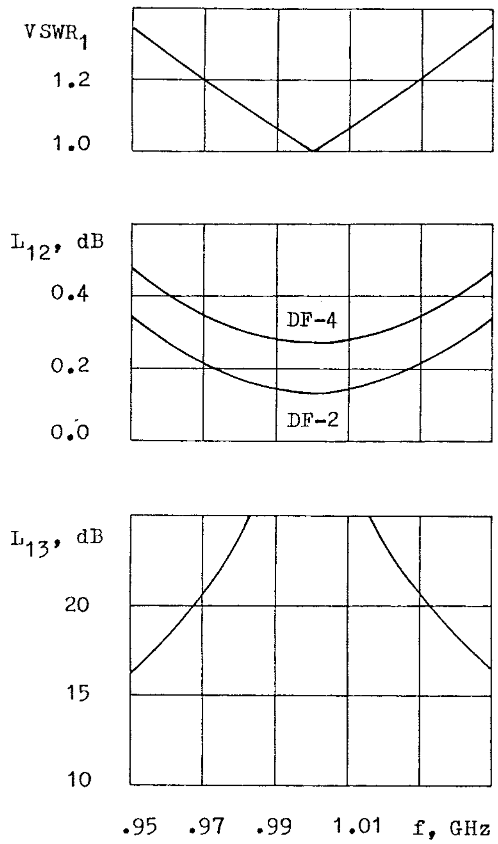


Fig. 5. Theoretical characteristics $VSWR_1(f)$, $L_{12}(f)$, and $L_{13}(f)$ evaluated for the filters DF-2 and DF-4 with forward-biased p-i-n diodes of type MA4P7000 (see Appendix).

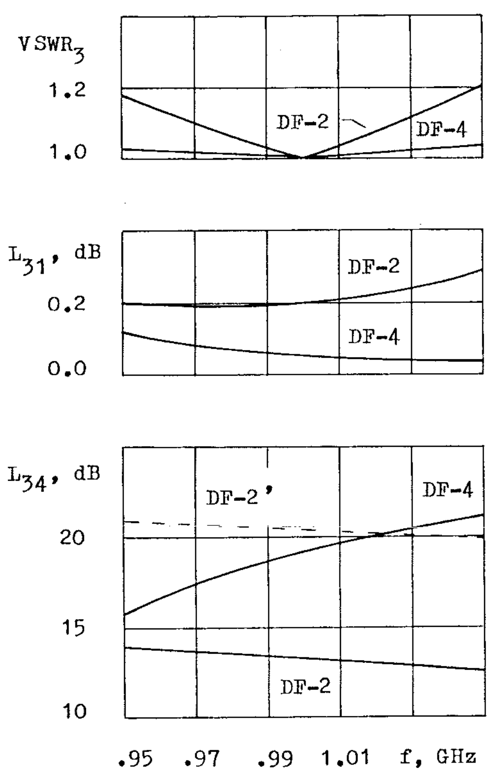
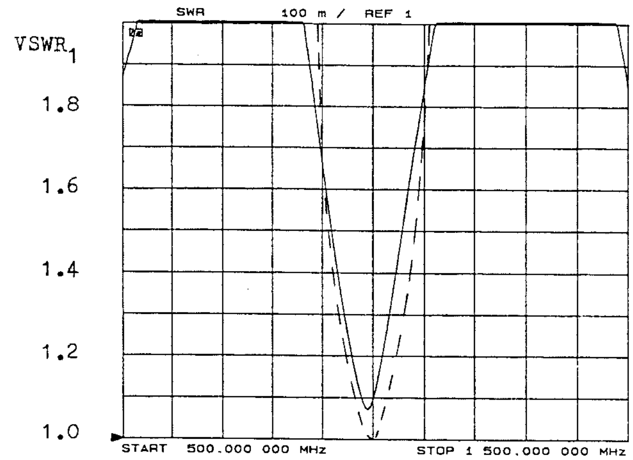
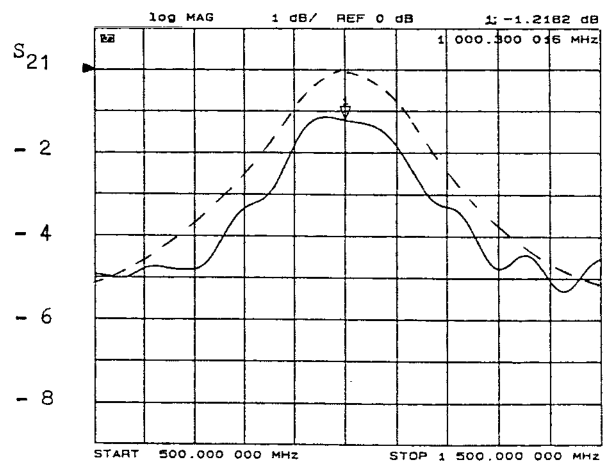


Fig. 6. Characteristics $VSWR_3(f)$, $L_{31}(f)$, and $L_{34}(f)$ calculated for the filters DF-2 and DF-4 with reverse-biased p-i-n diodes of type MA4P7000 (see Appendix).



(a)



(b)

Fig. 7. Characteristics (a) $VSWR_1$ and (b) S_{21} [dB] experimentally and theoretically (the dashed lines) obtained for the DF-4 filter with ideal p-i-n diodes in the on state simulated by zero-impedance switches. Similar characteristics are achieved for the DF-2 filter if $1/(\omega C_s) \cong 0$.

$X_c = 1/(\omega C_s)$ is negligible in comparison with the characteristic impedances Z and Z_0. The characteristics $L_{32}(f)$ and $L_{14}(f)$ are not illustrated because they are greater than 100 dB over a wide frequency band. If the diodes are reverse biased then characteristics $VSWR_3(f)$, $L_{31}(f)$, and $L_{34}(f)$ calculated for both filters are shown in Fig. 6. Also in this case, $L_{32}(f)$ is greater than 100 dB over the frequency range. The characteristic $L_{34}(f)$, shown by the dashed line, has been calculated for the filter DF-2 with a decreased value of the p-i-n diode capacitance C_t equal to 0.3 pF. It should be pointed out that the isolation characteristics $L_{14}(f)$ and $L_{32}(f)$ are independent of the parameters of the p-i-n diodes if the diodes are identical. All the characteristics presented have been calculated under the assumption that dissipative losses in the filter microstrip-line sections are negligible. Additionally, it has been assumed that the 180° phase-reversal units are ideal, i.e., that parameters L, C, and $1/Y_s$ of the equivalent circuit shown in Fig. 3 are equal to zero.

III. Experimental Results

In order to verify the theoretical predictions, the filters DF-2 and DF-4 (see Fig. 2) have been designed for: $Z_0 = 50\ \Omega$, $Z = 50\ \Omega$, $f_0 = 1$ GHz, $C_s = 2.2$ nF and ideal p-i-n diodes simulated by lossless electrical switches. The filters have been constructed in the microstrip-line technology using epoxy–glass substrates with

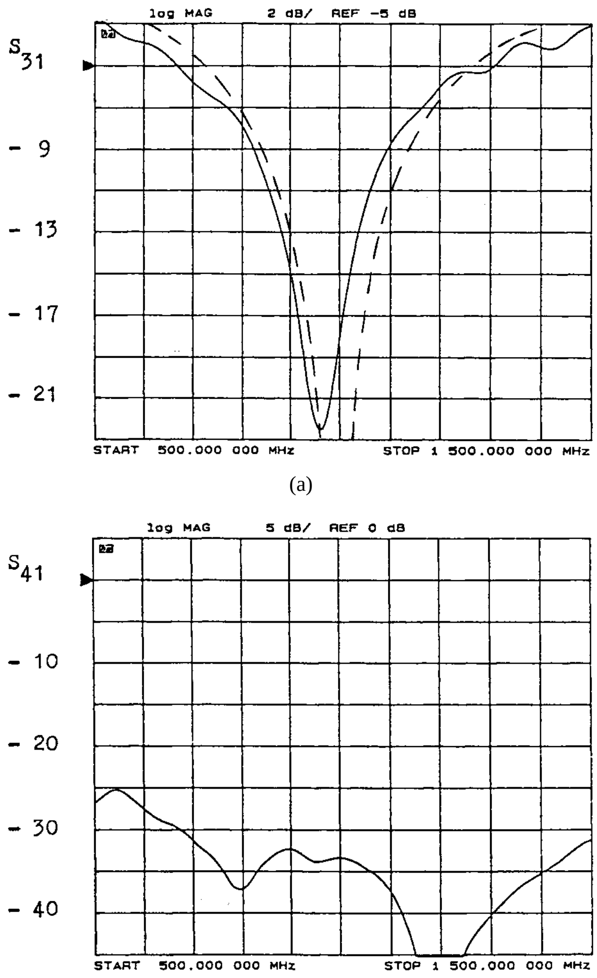


Fig. 8. Characteristics (a) S_{31} [dB] and (b) S_{41} [dB] experimentally and theoretically (the dashed lines) obtained for the DF-4 filter with ideal p-i-n diodes in the on state simulated by zero-impedance switches. The corresponding characteristics obtained for the DF-2 filter are almost the same if $1/(\omega C_s) \cong 0$.

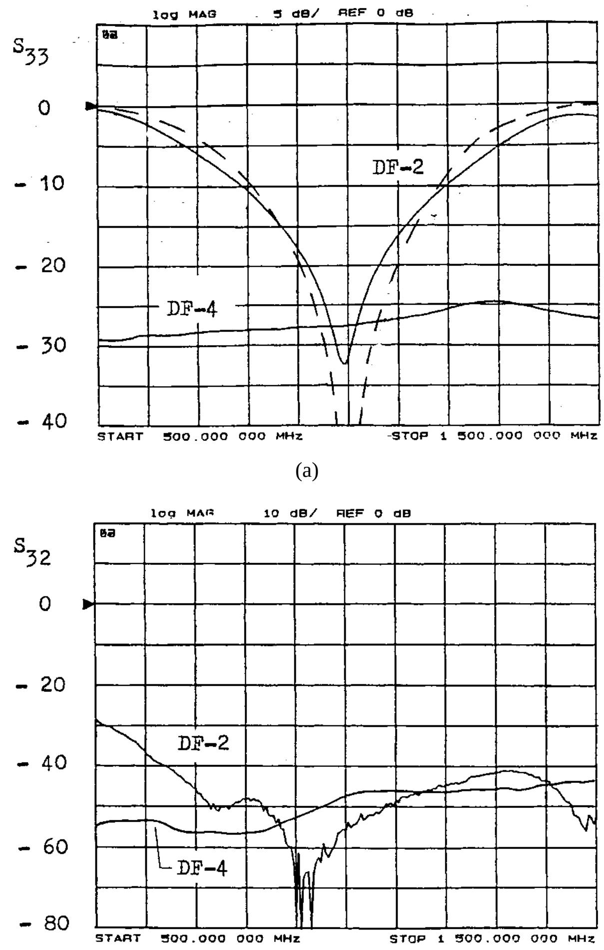


Fig. 9. Characteristics (a) S_{33} [dB] and (b) S_{32} [dB] experimentally and theoretically (the dashed lines) obtained for DF-2 and DF-4 filters with ideal p-i-n diodes in the off state. The switches simulating p-i-n diodes have been removed.

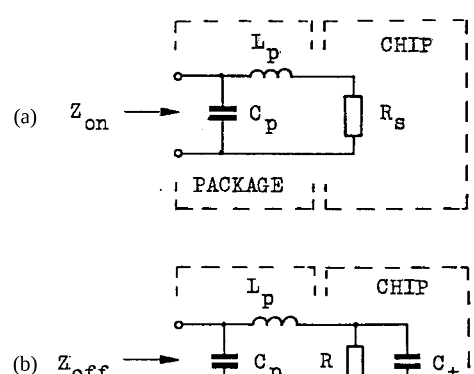


Fig. 10. Equivalent circuits for the p-i-n diodes under (a) forward bias and (b) reverse bias.

electrical permittivity $\epsilon_r = 4.23$, dielectric thickness h of 2 mm and strip thickness t equal to 0.035 mm. The geometry of the slot-line hollow patches has been evaluated experimentally [4], [5]. Figs. 7 and 8 present frequency characteristics $VSWR_1$, $S_{21}[\mathrm{dB}] = 20\log|\mathrm{S}_{21}|$, $S_{31}[\mathrm{dB}] = 20\log|\mathrm{S}_{31}|$ and $S_{41}[\mathrm{dB}] = 20\log|\mathrm{S}_{41}|$ obtained experimentally for both filters with the switches short-circuited. When the switches are in the off state then characteristics $S_{33}[\mathrm{dB}] = 20\log|\mathrm{S}_{33}|$ and $S_{32}[\mathrm{dB}] = 20\log|\mathrm{S}_{32}|$ are similar to that shown in Fig. 9. The corresponding theoretical characteristics are also illustrated in Figs. 7–9 by the dashed lines. It has been confirmed experimentally that characteristics $S_{41}(f)$ and $S_{32}(f)$ depend strongly on the construction of the phase-reversal unit. Its equivalent shunt admittances $Y(f) = j2\pi fC$ and Y_s due to the slot-line hollow patches [Figs. 3(b), (d)], should be as small as possible over a required frequency band. The admittance $Y(f)$ can be reduced by increasing the width W_s of the slot in which copper strips forming the 180° twist are placed. Similarly, by increasing the geometrical dimensions of the slot-line hollow patches, the characteristics $S_{41}(f)$ and $S_{32}(f)$ become more broad-band. They extend mainly toward the lower frequencies. From the analysis performed, it follows that the experimental results confirm well the theoretical predictions. As seen in Fig. 5, the four-diode filter (DF-4) has greater insertion loss than the two-diode one due to the additional p-i-n diodes. The filter DF-4, however, ensures better transmission between ports 3 and 1 (see Fig. 6). Moreover, its transmission characteristic $VSWR_3(f)$ is more broad-band in comparison with the corresponding one obtained for the filter DF-2 (Fig. 9).

IV. Conclusions

Two original electronically switchable microstrip-line directional filters are presented. They provide substantially improved frequency characteristics compared to the conventional constructions. The unique feature of them is that they offer high-level isolation, independent of frequency, over a wide frequency range for two pairs of opposite ports. These isolation characteristics are also independent of the p-i-n diode parameters if the diodes used are identical. It should be pointed out that the above conclusions relate to the filters operating in two modes (reception and transmission), i.e., when the diodes are forward or reverse biased. Thus, these directional filters seem to be suitable for some radio-communication applications.

Appendix

Fig. 10 presents equivalent circuits for MA4P7000 p-i-n diode under forward and reverse bias. The following parameters have been used for calculations: $C_p = 0.001$ pF, $L_p = 0.001$ nH, $R_s(I_0 = 100 \text{ mA}) = 0.8\ \Omega$, $C_T(U_R = 100 \text{ V}) = 0.70$ pF, and $R(U_R = 100 \text{ V}) = 200\ k\Omega$.

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Radiation Properties of a Planar Dielectric Waveguide Loaded with Conducting-Strip Diffraction Grating

Aleksander Błędowski and Władysław Żakowicz

***Abstract*—A dielectric planar waveguide periodically loaded with conducting strips is considered as a possible antenna for millimeter-wave range. We show that separating the grating from the waveguide leads to the reduction of radiative attenuation of the waveguided radiation and substantial narrowing of the angular spread of the diffracted radiation.**

***Index Terms*—Diffraction grating antennas, leaky-wave antennas.**

I. Introduction

Diffraction gratings placed on the surface of dielectric waveguides are very important components of many microwave devices serving either as transmitting or receiving antennas and also subject to numerous theoretical and experimental studies.

Our theoretical analysis, developed along a general electromagnetic theory of gratings [1] and [2], is devoted to a planar dielectric waveguide loaded with periodic infinitely thin metallic strips made of perfect conductor. We consider this system for a millimeter-wavelength radiation.

Similar systems have often been studied in the past in [3]–[6]. Generalizing these discussions, we admit an additional gap between the grating and the waveguide, allowing better control of the coupling between the guided and diffracted waves. For a weaker coupling and correspondingly slower attenuation, the outgoing beam can be concentrated in a much narrower angular sector. Often this is a desired beam property. Such a grating separation is present in a technical construction of a millimeter-wave antenna with variable grating period [8].

In Section II, the basic equations describing waves in the investigated structure are derived. We consider only the waves propagating perpendicularly to the strips and having either TE or TM polarizations. In Section III, numerical results showing the complex propagation constant as a function of the various waveguide and grating parameters are presented. Some discussion on the angular width of the diffracted radiation is also given.

II. Theory

We consider a dielectric waveguide of thickness h and dielectric constant ϵ_w loaded with a diffraction grating formed by parallel metallic strips distributed periodically with a period d. The grating is separated from the waveguide by a dielectric layer of thickness b and dielectric constant ϵ_b. The metallic strips are made of perfect conductor, have width w, and infinitesimal thickness. The whole system is surrounded by a homogeneous dielectric medium with dielectric constant ϵ_v. The system, along with a coordinate system used, are shown in Fig. 1.

We assume the electromagnetic waves to be monochromatic with the time-dependence factor $\exp(i\omega t)$, perpendicularly propagating

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