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SWITCH-FILTER ON A RECTANGULAR WAVEGUIDE PARTIALLY FILLED BY DIELECTRIC

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Abstract. *At article investigates a broadband switch for microwave technology, antenna-feeder paths of which are implemented on a rectangular waveguide partially filled by dielectric. Modern means of communication of the microwave range can operate for transmission through two independent antenna channels, each of which includes its own microwave transmitter. It is also provided the operation of one microwave transmitter for two antennas. The transmission of high-power signals requires the implementation devices based on rectangular waveguide partially filled by dielectric. The active element is an open nonlinear structure included in a dielectric plate which located in a rectangular waveguide. The electrodynamic problem is solved by the method of eigenfunctions. In this article, the transfer coefficient and plotted the graphs of the dependence of the electrical length of the waveguide segment with the open nonlinear structure on the value of the reactive conductivity of the inductive loop at fixed reactive conductivity of the open nonlinear structure is determine. The results of this article can be used in the development of broadband switches for mobile digital combined troposcatter-radio relay stations with space-diversity transmission, antenna-feeder paths of which are implemented on the rectangular waveguide partially filled by dielectric. The article developed a filter on waveguide tees partially filled by dielectric. A design based on an H-tee on a rectangular waveguide partially filled by dielectric is proposed. The equivalent circuit of the H-tee is based on the equivalent circuit in the longitudinal slot in the narrow wall of the waveguide. Such a gap interrupts the lines of the transverse surface current of the main wave of a rectangular waveguide partially filled by dielectric. Numerical results are obtained for transformation ratios.*

Keywords: rectangular waveguide partially filled by dielectric, switch, filter, open nonlinear structure

PRZEŁĄCZNIK-FILTR NA FALOWODZIE PROSTOKĄTNYM CZĘŚCIOWO WYPEŁNIONYM DIELEKTRYKIEM

Streszczenie. *W artykule zbadano szerokopasmowy przełącznik dla techniki mikrofalowej, którego tory antenowo-podajnikowe są realizowane na falowodzie prostokątnym częściowo wypełnionym dielektrykiem. Nowoczesne urządzenia do komunikacji mikrofalowej mogą pracować na dwóch niezależnych kanałach antenowych, z których każdy zawiera własny nadajnik mikrofalowy. Możliwa jest również praca jednego nadajnika mikrofalowego na dwóch antenach. Transmisja sygnałów dużej mocy wymaga realizacji urządzeń opartych na falowodzie prostokątnym częściowo wypełnionym dielektrykiem. Elementem aktywnym jest otwarta struktura nieliniowa zawarta w płytce dielektrycznej, która znajduje się w prostokątnym falowodzie. Problem elektrodynamiczny jest rozwiązywany metodą funkcji własnych. W artykule wyznaczono współczynnik transmisji oraz wykreślono wykresy zależności długości elektrycznej segmentu falowodu o otwartej strukturze nieliniowej od wartości reaktancji obwodu indukcyjnego przy ustalonej reaktancji otwartej struktury nieliniowej. Wyniki artykułu mogą być wykorzystane przy opracowywaniu szerokopasmowych przełączników ruchomych cyfrowych kombinowanych stacji przekaźnikowych troposferycznych z separacją przestrzenną, których tory antenowo-podajnikowe realizowane są na falowodzie prostokątnym częściowo wypełnionym dielektrykiem. W artykule opracowano filtr na trójnikach falowodowych częściowo wypełnionych dielektrykiem. Zaproponowano konstrukcję opartą na trójniku typu H na falowodzie prostokątnym częściowo wypełnionym dielektrykiem. Obwód równoważny trójkąta H jest oparty na obwodzie równoważnym w szczelinie podłużnej w wąskiej ściance falowodu. Szczelina taka przerywa poprzeczne linie prądu powierzchniowego fali fundamentalnej w falowodzie prostokątnym częściowo wypełnionym dielektrykiem. Otrzymano wyniki numeryczne dla współczynników transformacji.*

Słowa kluczowe: prostokątny falowód częściowo wypełniony dielektrykiem, przełącznik, filtr, otwarta struktura nieliniowa

Introduction

Combined radio engineering systems are one of the trends in the development of microwave technology [1-4, 8, 9]. In article [7], a mobile digital tropospheric radio relay station is analyzed, which includes two microwave transmitters. This station can transmit on two independent antenna channels, each of which includes its own microwave transmitter. It is possible to operate one microwave transmitter on two antennas; the second microwave transmitter is on standby. Different modes of operation of the microwave transmission path are equipped with two on-off microwave switches [7]. The requirement to ensure broadband operation of the microwave transmission path can be met by implementing the path on a rectangular waveguide partially filled by dielectric (RWPFD). In this case, it is necessary to implement two-position microwave switches also on RWPFD.

The aim of the work is to develop a broadband microwave switch on the RWPFD and a trap filter on the RWPFD.

1. Microwave switch on RWPFD

The microwave switch incorporates a packageless diode, which is included in a dielectric plate in the form of an open nonlinear structure (ONS). The inclusion of ONS can be both parallel and serial.

With the serial connection of the ONS in the RWPFD, a wide bandwidth can be obtained, but the capacitance of the ONS should be minimal. Parallel connection of ONS allows to provide a higher level of switching power than with series connection. It should be noted that it is the implementation of the switch on the RWPFD that provides a wider operating frequency band and a higher dielectric strength [6]. In addition, it is possible to expand the operating frequency band if the parallel connection of the ONS in the RWPFD is carried out together with a loop that compensates for the capacitance of the ONS.

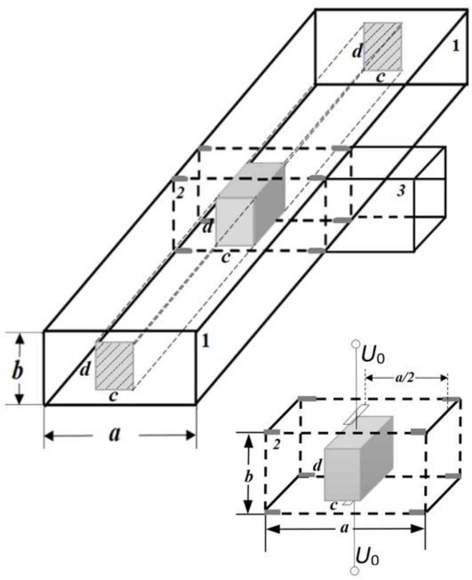

Fig. 1. The design of the switch on RWPFD with a compensating loop

On Fig. 1 shows the design of the switch on the RWPFD with a compensating loop. This RWPFD is a rectangular waveguide with dimensions $a \times b$, in which a dielectric plate with dimensions $c \times d$ is located, which does not affect the walls of the waveguide and has a relative permittivity ε_r. A ONS of the same geometric dimensions is included in the dielectric plate. Waveguide 1 is a RWPFD, waveguide 2 is a RWPFD with ONS, waveguide 3 is a hollow rectangular waveguide.

The ONS is controlled through a non-radiating longitudinal slot in the wide wall of the RWPFD. The value of the resistance of the ONS is a few ohms in the presence of a control voltage and thousands of ohms in its absence. To compensate for the capacitance of the ONS, which should be as small as possible, a compensating inductive loop is turned on in the form of a segment of a hollow rectangular waveguide from the side of the narrow wall of the main waveguide. Such an inclusion is an *H*-connection of waveguides.

On Fig. 2 shows the equivalent circuit of the plane-transverse joint of the RWPFD with the ONS, the plane-transverse joint of the ONS with the RWPFD and a parallel-connected compensating loop. On Fig. 2 shows the normalized reactive conductivity jb_c of the plane-transverse junction of the RWPFD with ONS and the ONS with RWPFD, the transformation coefficients N_0 for the main quasi-H_{10} wave and the normalized reactive conductivity $-jb_L$ of the compensating inductive loop.

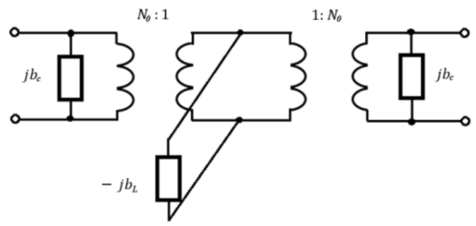

Fig. 2. Equivalent circuit of the plane-transverse joint of the RWPFD with ONS, the plane-transverse joint of the ONS with RWPFD and a parallel-connected compensating loop

"Collapsing" the circuit shown on Fig. 2 through the recalculation of the transformation ratios N_0, we obtain an equivalent circuit with parallel normalized conductivity jb_Σ and normalized reactive conductivity $-jb_L$ of the compensating inductive loop (Fig. 3).

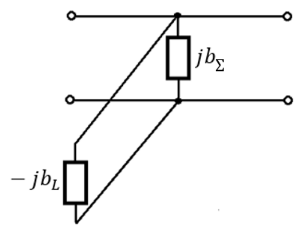

Fig. 3. Equivalent circuit with parallel normalized conductivity jb_Σ and normalized reactive conductivity $-jb_L$ of a compensating inductive loop

Calculation of the normalized reactive conductivities was carried out according to the following formulas:

$$b_c = \frac{1}{N_0^2}\sum_{k=1}^{\infty} N_k^2 y_k \tag{1}$$

$$b_\Sigma = \frac{2b_c}{N_0^2 y_0} \tag{2}$$

$$N_0 = \int_S \bar{\mathcal{E}}_{h_{10}} \bar{\mathcal{E}}_H \, dS$$

$$N_k = \int_S \bar{\mathcal{E}}_{h_{10}} \bar{\mathcal{E}}_k dS$$

$$\bar{\mathcal{E}}_{h_{10}} = \sqrt{128/ab\left(64+q^2+p^2+q^2p^2\right)}\,\frac{1}{k_{h_{10}}} * \mathcal{F} \tag{3}$$

$$\mathcal{F} = \left\{\left[\left(\frac{\pi}{a}\right)sin\frac{\pi x}{a} - \left(\frac{p\pi}{2a}\right)sin\frac{\pi x}{a}cos\frac{2\pi y}{b} - \left(\frac{3q\pi}{ba}\right)* \right.\right.$$
$$*sin\frac{3\pi x}{a}sin\frac{\pi x}{a}cos\frac{2\pi y}{b} - \left(\frac{3q\pi}{ba}\right)sin\frac{3\pi x}{a} +$$
$$\left. + \left(\frac{3qp\pi}{16a}\right)sin\frac{3\pi x}{a}cos\frac{2\pi y}{b}\right]\bar{y}^0 + \left[\left(\frac{p\pi}{b}\right)cos\frac{\pi x}{a} * \right.$$
$$\left.\left. *sin\frac{2\pi y}{b} - \left(\frac{2qp\pi}{8b}\right)cos\frac{3\pi x}{a}sin\frac{2\pi y}{b}\right]\bar{x}^0\right\}$$

$$\bar{\mathcal{E}}_H = \sqrt{128/ab\left(64+q^2+p^2+q^2p^2\right)}\,\frac{1}{k_H} * \mathcal{F} \tag{4}$$

$$k_H^2 = k_0^2\left(\varepsilon_{ef}\left(1+\alpha\bar{\mathcal{E}}_{10}^2\right)\right) - \gamma_{10}^2$$

where $\bar{\mathcal{E}}_k$ – transverse electric eigenfunction of higher types of quasi-H_{30} and quasi-H_{50} waves RWPFD [6]; $\bar{\mathcal{E}}_{10}$ – transverse electric eigenfunction of a hollow rectangular waveguide [6]; k_0, γ_{10} – the wavenumber of a hollow rectangular waveguide and the longitudinal wavenumber of the RWPFD respectively [6]; ε_{ef} – effective permittivity [6]; y_0, y_k – the normalized conductivity of the main quasi-H_{10} wave and the normalized conductivity of higher types of quasi-H_{30} and quasi-H_{50} waves RWPFD, respectively [6].

As a local field, in the first approximation, we take into account the quasi-H_{30} and quasi-H_{50} waves:

$$\bar{\mathcal{E}}_{h30} = \sqrt{128/ab\left(64+q^2+p^2+q^2p^2\right)} * \frac{1}{k_{h30}} * \mathcal{J} \tag{5}$$

$$\mathcal{J} = \left\{\left[\frac{p\pi}{b}cos\frac{3\pi x}{a}sin\frac{2\pi y}{b} - \frac{qp\pi}{16b}\left(cos\frac{5\pi x}{a} - 2cos\frac{\pi x}{a}\right)* \right.\right.$$
$$\left. *sin\frac{2\pi y}{b}\right]\bar{x}^0 + \left[\frac{3\pi}{a}sin\frac{3\pi x}{a} - \frac{3p\pi}{2a}sin\frac{3\pi x}{a}cos\frac{2\pi y}{b} - \right.$$
$$-\frac{q\pi}{8a}\left(\frac{5}{2}sin\frac{5\pi x}{a} - sin\frac{\pi x}{a}\right) +$$
$$\left.\left. +\frac{qp\pi}{16a}\left(\frac{5}{2}sin\frac{5\pi x}{a} - sin\frac{\pi x}{a}\right)cos\frac{2\pi y}{b}\right]\bar{y}^0\right\}$$

$$\bar{\mathcal{E}}_{h_{50}} = \sqrt{128/ab\left(64+q^2+p^2+q^2p^2\right)} * \frac{1}{k_{h_{50}}} * K \tag{6}$$

$$K = \left[\left(\frac{p\pi}{b}\right)cos\frac{5\pi x}{a}sin\frac{2\pi y}{b} - \left(\frac{qp\pi}{8b}\right)K_1 sin\frac{2\pi y}{b}\right]\bar{x}^0 +$$
$$+\left[\left(\frac{5\pi}{a}\right)sin\frac{5\pi x}{a} + \left(\frac{qp\pi}{16a}\right)K_2 cos\frac{2\pi y}{b} - \right.$$
$$\left. -\frac{qp\pi}{8a}K_2 - \left(\frac{5p\pi}{2a}sin\frac{5\pi}{a} - cos\frac{2\pi y}{b}\right)\right]\bar{y}^0$$

$$K_1 = \frac{1}{3}cos\frac{7\pi x}{a} - \frac{1}{2}cos\frac{3\pi x}{a}$$

$$K_2 = \frac{7}{3}sin\frac{7\pi x}{a} - \frac{3}{2}sin\frac{3\pi x}{a}$$

Note that the dielectric plate, in contrast to the ONS, which is an isotropic nonlinear dielectric from ε_d, is an isotropic linear dielectric with ε_r. Even with the passage of current through the ONS, the nonlinear process will proceed in a weak electromagnetic field. Therefore, the dependence of the electrical induction on the electrical intensity will be linear in the working range of changes in the magnitude of the electrical intensity. In accordance with [5], this is a local linearization of the above dependence. Taking into account that the electromagnetic field will also change slightly in the local hollow area of the RWPFD, located in the immediate vicinity of the ONS ($\mathscr{L} << \Lambda_{h_{10}}$, where $\mathscr{L}$ is the maximum size of the ONS, $\Lambda_{h_{10}}$ is the length of the main quasi-H_{10} wave of the RWPFD), the expression for the relative electric ONS permeability will have following:

$$\varepsilon_d = \left(1+\alpha\bar{\mathcal{E}}_{10}^2\right) \tag{7}$$

where α – coefficient of nonlinearity, depending on the material of the ONS [5].

The transmission coefficient T_{11} for the circuit in Fig. 3 has the following form:

$$T_{11} = 1+\frac{1}{2}\left[jb_\Sigma\Delta f - jb_L ctg\left(\theta\Delta f\right)\right] \tag{8}$$

where b_Σ, b_L – normalized reactive conductivities of the RWPFD with ONS and inductive loop respectively, $\Delta f = f/f_o$ – frequency range, θ electrical length of the inductive loop. At the resonant frequency $T_{11} = 1$, then we have:

$$\theta = arcctg\frac{b_\Sigma}{b_L} \tag{9}$$

On Fig. 4 shows the dependence of the value of θ on the value of b_L at fixed b_Σ: curve 1 is plotted at $b_\Sigma = 0.5$, curve 2 – at $b_\Sigma = 1$ These graphs can be used when calculating switches in transmission modes (de-energized ONS).

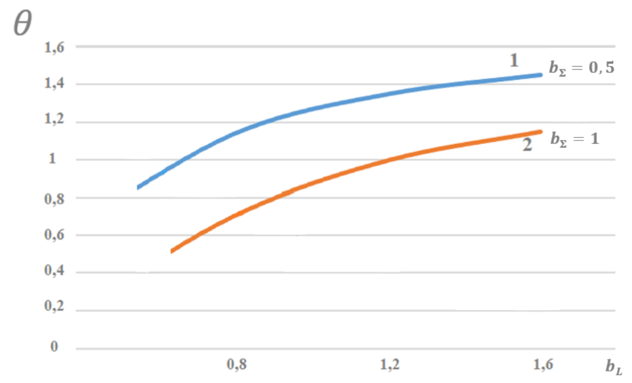

Fig. 4. Dependence of the value of θ on the value of b_L at fixed b_Σ

2. Filter on RWPFD

If a diaphragm is added to the design of Fig. 1 at the junction of the loop with the main waveguide and the loop is short-circuited with a short-circuiting plate.

The design of the filter on the *H*-tee is shown in Fig. 5.

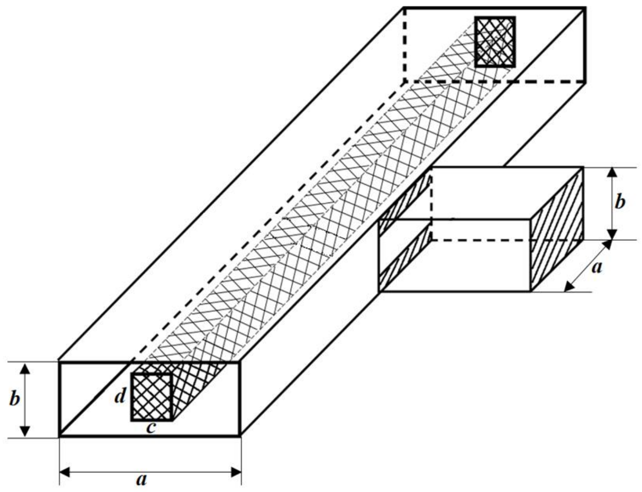

Fig. 5. Filter design on H-tee

The equivalent circuit of the filter on the *H*-tee is shown in Fig. 6.

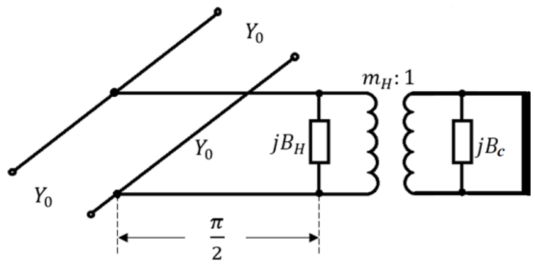

Fig. 6. Equivalent circuit for connecting the main waveguide with a side waveguide in an H-tee with a piston

When an electromagnetic wave falls from the *H*-arm of the tee, the electric fields in the side arms at equal distances from the joint are in phase, and the magnetic fields are in antiphase. The equivalent circuit of the *H*-tee is based on the equivalent circuit of the longitudinal slot in the narrow wall of the waveguide. Such a slit interrupts the lines of the transverse surface current of the main quasi-H_{10} wave of the RWPFD. The reaction of local fields in the vicinity of the slot is displayed by reactive conductivity jB_H and the conversion of the field amplitude at the slot to the amplitude of the propagating wave is shown by switching on an ideal transformer with a transformation coefficient m_H.

In order for the circuit to be valid under any load, including short circuit, the metallization of the slot must turn the waveguide into a regular one. This can be achieved by turning on a quarter-wave section of the transmission line having an electrical length of $\pi/2$ at all frequencies. The short-circuited quarter-wave transmission line does not affect the passage of the wave in the main waveguide.

Conductivity B_C is determined as follows:

$$B_C = \sum_k n_k Y_k$$
$$n_k = \int_S \bar{\mathcal{E}}_{h_{10}} \bar{\mathcal{E}}_k dS \tag{10}$$

The expressions for the transformation coefficients and reactive conductivities are as follows:

$$m_H = \int_S \bar{Э}_H \bar{\mathcal{E}}_{h_{10}} dS$$
$$B_H = -\sum_k Y_k \left[\left(k_k^2+\gamma_k^2\right)/\gamma_k^2\right] \quad \int_S \bar{Э}_H \bar{\mathcal{E}}_k dS$$

where k_k – transverse wavenumbers of higher wave types RWPFD, γ_k – longitudinal wavenumbers of higher wave types RWPFD [6]. The coordinate function $\bar{Э}_H$ [7] approximating the field distribution at the hole has the following form:

$$\bar{Э}_H = \sqrt{128/\left(64+q^2\right)}\ Ce_1\left(x,q\right)\bar{z}^0$$

where Ce_1 – even Mathieu function of the first kind, q – fill parameter of RWPFD.

Also note that in a frequency-tunable filter, the side arm is a segment of a hollow rectangular waveguide, as shown in Fig. 5. In a filter that does not tune in frequency, the side arm is RWPFD.

Tables 1 show the values of the transformation coefficient m_H for formule (11) depending on the value of $\frac{2a}{\Lambda}$.

Table 1. Dependence of the transformation coefficients on the normalized value

m_H	1.7	1.5	1.3	0.91	0.75	0.69	0.61	0.53
$\frac{2a}{\Lambda}$	0.7	0.8	0.9	1.1	1.2	1.3	1.4	1.5

3. Conclusions

In conclusion, we note that the developed broadband switch on the RWPFD can be used for microwave transmission paths of mobile digital troposcatter stations, mobile space communication stations and mobile digital combined troposcatter-radiorelay stations, which include two microwave transmitters. The prospects for the use of such broadband switches are significantly increased due to the introduction of space-diversity transmission into microwave technology. Once again, we note that the broadband bandwidth of the device was achieved not only due to design features, but also due to the implementation of the design on the RWPFD.

The inclusion of an open-frame semiconductor diode in the form of an ONS allows the device to function as a two-position switch: in the presence of a control current, the resistance of the ONS is units Ohm, in the absence – thousands Ohms. Such a switch on RWPFD provides two-position switching of high transmitting power in a wide frequency band. The high dielectric strength of the RWPFD was investigated in [6].

When implementing a filter on rectangular waveguides with dimensions of 40×20 mm^2 with a dielectric plate with $\varepsilon_r = 4$, it is possible to ensure the operation of the mobile digital troposcatter radio relay station radio relay component in two frequency ranges 3.8...4.2 GHz + 5.925...6.425 GHz [10, 11] with one antenna-feeder path. This filter provides a blockband of 4.2...5.925 GHz.

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