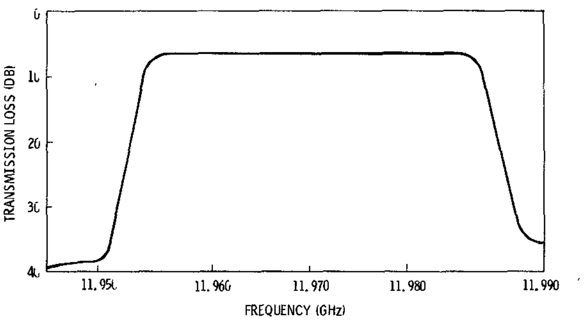

Fig. 7. Measured transmission loss response for an 8-pole elliptic function filter with a 4-pole equalizer.

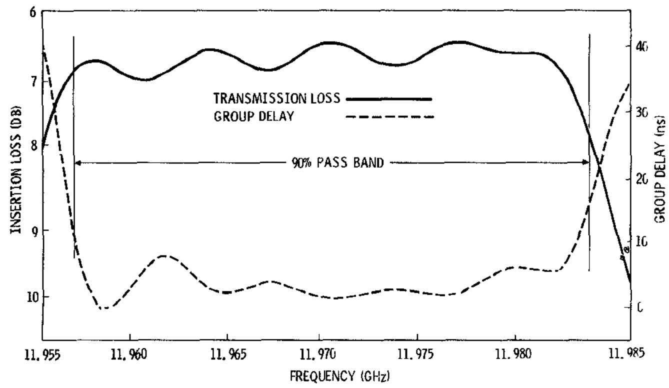

Fig 8. Measured inband insertion loss and group delay responses for an 8-pole elliptic function filter with a 4-pole equalizer.

V. Conclusion

A simplified method for equalizer design has been developed. This method uses a commonly available computer subroutine to solve for the equalizer parameters directly instead of the conventional way of solving for the equalizer poles. [2]. Thus, the step of realization, which generates the equalizer parameters from the equalizer poles, is eliminated. This method allows the equalizer to be constructed with a larger number of poles. Therefore, a single equalizer would be able to perform a larger degree of equalization. Consequently, a considerable weight reduction can be obtained over the conventional approach of using cascaded equalizers.

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Improved Selectivity in Cylindrical TE_{011} Filters by TE_{211}/TE_{311} Mode Control

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***Abstract*—A new method is presented for the design of low loss cylindrical TE_{011}-mode resonators whereby transmission nulls can be placed near the TE_{011} resonance by controlling the TE_{211} and TE_{311} modes that are naturally excited in the same resonator. The frequencies at which the nulls occur are controlled by the angular offset of the sidewall coupling apertures and the relative amplitude of the TE_{011} mode compared to the TE_{211} and TE_{311} modes. It is also shown that a lumped constant circuit model can be used to accurately represent the multimode response of the resonator.**

I. Introduction

The high unloaded Q of the cylindrical TE_{011} mode is attractive for low loss filters, especially at the higher microwave frequencies where transmitter power and receiver sensitivity are often limited and expensive. The design of cylindrical TE_{011}-mode filters is complicated, however, by the large number of modes that resonate at frequencies close to or degenerate with the TE_{011} mode. The response of these modes must be controlled to obtain usable filter characteristics. The TM_{111} mode is of particular concern because it is degenerate with the TE_{011} mode in the right cylindrical resonator. The TE_{112}, TE_{211}, TE_{311}, TM_{011}, TM_{012}, TM_{110}, and TM_{210} modes can also seriously affect the filter performance depending on the particular application and the filter design. The presence of these modes also makes it difficult to compute the filter response by the techniques usually used. TE_{011}-filter design is thus primarily an experimental problem.

The relative frequencies of the resonances of the TE_{011} and the other modes, with the exception of the degenerate TM_{111}, can be controlled, within limits, by the choice of the diameter-to-length ratio of the cavity [1]. Large changes in the diameter-to-length ratio can result, however, in significant reduction in the unloaded Q. Atia and Williams [2] have shown that the TM_{111} resonant frequency can be separated from the TE_{011} resonance by thin metal posts or by dielectric material on the cavity end walls. Thal [3] has shown that a similar effect can be obtained from shaping the cavity by chamfering the edges. The degree of shaping can also be used to control the relative frequencies of the modes without degrading the unloaded Q of the TE_{011} mode. Cavity shaping is particularly attractive because it permits all the resonators of a filter to have different shapes, so that when they are

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synchronously tuned to the TE_{011} mode they are not synchronously tuned for the other modes.

Suppression and control of the unwanted modes is still necessary, though, even with cavity shaping. TM modes can be suppressed by using narrow rectangular apertures that are oriented to favor coupling to the TE modes. The coupling through the apertures is approximately proportional to the cube of its length in the direction of the magnetic field, so the reduction in the TM-mode response can be large. The TE_{112} mode can be suppressed by using sidewall coupling at the center of the resonator since the mode possesses odd symmetry about the center. The most difficult modes to control are the TE_{211} and TE_{311} modes. Atia and Williams [2] achieved excellent results in suppressing these modes, apparently aided in part by using a combination of endwall and sidewall coupling. The use of endwall coupling is, however, not always possible because of configuration and other constraints. Thal [3] and others [1] attempted to control the TE_{211} and TE_{311} modes in sidewall-coupled cavities by using angular offsets between the coupling apertures rather than having the apertures directly opposite one another in the resonator. This permits one aperture to be at a location of low field strength for the unwanted modes excited by the other aperture. This has been of limited success as it usually results in asymmetrical rejection characteristics due to TE_{211}/TE_{311} mode contamination.

It is the purpose of this paper to show that even if TE_{211} and TE_{311} modes cannot be suppressed in sidewall-coupled cavities, the modes can be controlled in a manner that can enhance the filter performance for many applications. In addition, a lumped constant circuit representation of the cylindrical resonator is presented which accurately describes the response of the resonator over a large frequency range. This representation includes not only the TE_{011} mode but also the TE_{211} and TE_{311} modes.

II. TE_{211}/TE_{311} Mode Control Considerations

The coupling to a cylindrical resonator is solely magnetic for the TE_{011} mode and predominantly magnetic for the TE_{211} and TE_{311} modes. It is easier, however, to visualize the fields at the center of the cavity in terms of electric fields rather than magnetic fields. Fig. 1 shows the electric fields oriented for excitation by an input aperture with magnetic coupling. As can be seen in Fig. 1, there is no angular variation of the TE_{011} electric fields, so coupling through the resonator should not be a function of the angular relationship of the input and output coupling apertures. The coupling through the TE_{211} and TE_{311} modes, however, will vary with the angular offset of the coupling apertures. Minimum magnetic coupling occurs at offset angles of 45° and 135° for the TE_{211} mode and at angles of 30°, 90°, and 150° for the TE_{311} mode. There is no angular offset that simultaneously minimizes coupling to both modes. An angular offset in the range of 135°–150°, however, could provide a reasonable compromise for reduced coupling to both modes.

The angular offset of the apertures affects not only the magnitude of the coupling but also the relative phase. Fig. 1 shows that for the angle θ the response for the TE_{011} and TE_{211} modes would be in phase at the output port but the TE_{011} and TE_{311} modes would be out of phase. Other angular relationships permit these phase relationships to be reversed, or for the TE_{011} to be simultaneously in phase or out of phase with both the TE_{211} and TE_{311} modes. The angular offset of the coupling apertures can thus control not only the relative amplitude of the response but also the relative phase.

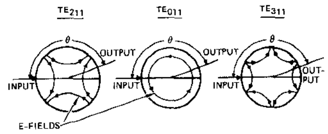


Fig. 1. Electric field configurations of modes.

The magnetic coupling of the apertures is proportional to the wall currents at the coupling aperture. The low loss of the TE_{011} mode is primarily due to the low sidewall currents. The TE_{211} and TE_{311} modes, however, have strong sidewall currents that are in the same direction as the currents associated with the TE_{011} mode. For a given sidewall coupling aperture, the coupling is thus much stronger for the TE_{211} and TE_{311} modes than the TE_{011} mode. This is the phenomenon that makes the design of sidewall-coupled TE_{011} mode filters, of any appreciable bandwidth, very difficult. Increasing the coupling to the TE_{011} mode also increases the already strong coupling to the TE_{211} and TE_{311} modes. The strong coupling reduces the resonant frequency, thus moving the TE_{311} resonance closer to the TE_{011} resonance, and it also reduces the loaded Q.

III. Experimental Data on Single Cavity Resonators

Because of the complexity of the mode interaction problems, experimental data was needed. Three single cavities (cylindrical, shaped, and sidewall-coupled) were constructed and tested to obtain data on the effect of variation of the angular offset of the coupling apertures. The cavities used shaping, similar to the "B" shape described by Thal [3], and had a moveable endwall on one end to adjust the resonant frequency. The cavity diameter was 0.500 in. The coupling apertures were rectangular in shape (0.114 in by 0.053 in), with the long dimension parallel to the axis of the cylindrical cavity. The three cavities were identical with the exception of the angular offsets of the coupling apertures. Angular offsets of 130°, 140°, and 150° were used. The three cavities are shown in Fig. 2.

The insertion loss measurements, made on the three cavities with a Hewlett-Packard R8747A K_a band transmission and reflection test unit, are shown in Fig 3. As expected, the response of the TE_{011} mode, shown in the center of the frequency range, does not vary with changes in the angular offset of the coupling apertures. There are, however, large changes in the response of the TE_{211} and the TE_{311} modes shown, respectively, on the lower and the upper ends of the frequency range. Also visible in Fig. 3 is a small response just below the frequency of the TE_{311} resonance. The response is the TE_{112} mode which is not fully suppressed. This is believed to be due to asymmetries introduced by the moveable endwall tuner.

The responses of the TE_{211} and TE_{311} modes are not those of single resonators. Each resonance appears to be that of two resonators which are overcoupled and whose coupling coefficient is a function of the angular offset of the coupling apertures. The response is similar to that of two coupled orthogonal modes in a cylindrical cavity as shown in Ragan [4].

The nulls that appear between the TE_{011} resonance and the TE_{211} and TE_{311} resonances are of even greater interest. The frequency of the nulls and even the existence of a null is a function of the angular offset of the apertures. Although not immediately obvious in the plots, the skirts of the TE_{011} response

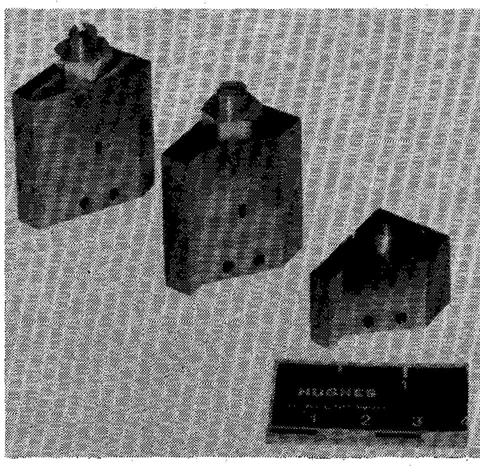


Fig. 2. Single-cavity resonators with various angular offsets of coupling apertures.

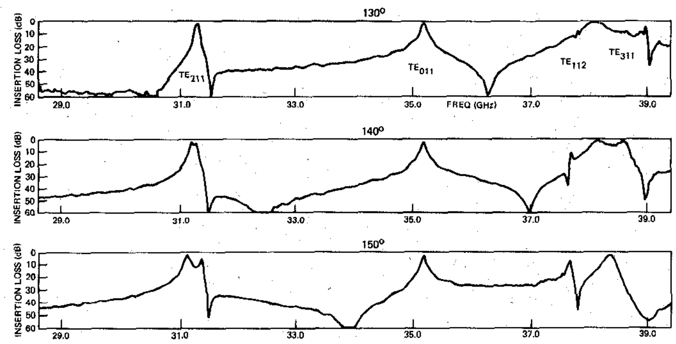


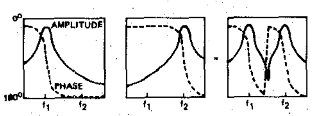
Fig. 3. Measured insertion loss of single-cavity resonators with various aperture offset angles.

are steepened by the presence of a null that is close in frequency to the TE_{011} response. The closer the null is, the steeper the skirts become.

IV. Transmission Nulls Due to Mode Interaction

The most important characteristic shown in the experimental data is the existence of transmission minima, or nulls, at frequencies between the TE_{211}-, TE_{011}-, and TE_{311}-mode resonances. The existence and frequency of these nulls are functions of the angular offset of the coupling apertures. The reason for this is readily apparent when one considers two facts: 1) the field at the coupling aperture is the vector sum of the components from all of the modes; and 2) there is a 180° shift in the transmission phase of a resonant circuit in passing through the resonant frequency.

Consider the response of a single circuit, or mode, that is resonant at frequency f_1, as shown in Fig. 4(a). The circuit has an amplitude response shown by the solid line and a phase response shown by the dotted line. The resonator has a small, but significant, response at frequency f_2, but the phase has changed by almost 180° in passing through the resonance. The response of a second identical circuit, but resonant at f_2, is shown in Fig. 4(b). If both circuits are modeled as series resonant and connected in parallel, the response of the combined circuit will have a null between f_1 and f_2, as shown in Fig. 4(c). The null is due to the fact that the amplitudes are identical at that frequency but there is a 180° difference in phase, thus creating a null. This case is



(a) (b) (c)

Fig. 4. Response of two interacting modes. (a) First mode. (b) Second mode. (c) Sum of mode responses.

similar to that between the TE_{011} and TE_{211} modes for the angle θ as shown in Fig. 1. As depicted, the fields at the output aperture are in phase according to the mode configuration. At frequencies between the TE_{211} and TE_{011} resonances, these fields associated with the two modes are actually out of phase because one is above resonance and the other is below resonance.

Refer again to Fig. 1 and the relationship of the TE_{011} and TE_{311} modes. In this case the fields at the output apertures are depicted as out of phase. At frequencies between these two resonances, however, one mode is above resonance and the other is below, thus giving an additional 180° phase shift. The total phase shift is thus 360°, so the fields are in phase and there is no null.

The above discussion is for single resonant cavities or circuits. The measured responses, however, of the TE_{211} and TE_{311} modes, shown in Fig. 3, appear as two overcoupled resonant circuits with the coupling a function of the angular relationship of the coupling apertures. The two resonators would have a phase shift of 360° rather than 180°. The 360° phase shift is not consistent with

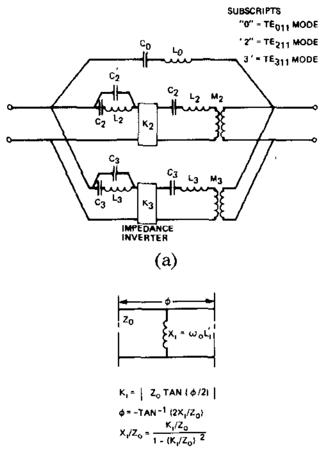


Fig. 5. Lumped constant circuit model. (a) Single-cavity representation. (b) Impedance inverter model and design equations.

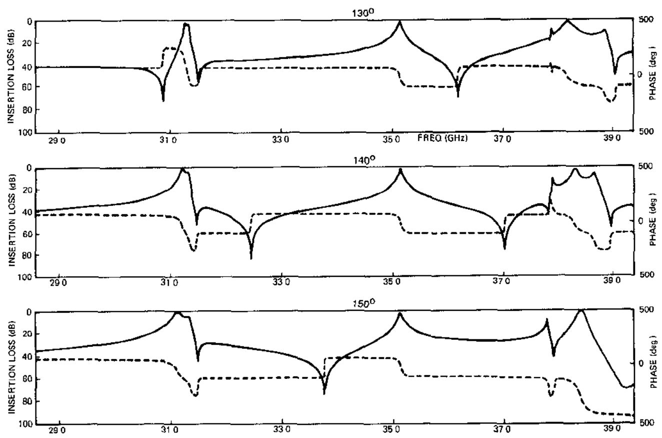


Fig. 6. Calculated transmission characteristics of single-cavity resonators with various aperture offset angles.

TABLE I

CAPACITANCE IN FARADS, INDUCTANCE IN HENRIES, AND RESISTANCE IN OHMS: LOAD AND SOURCE IMPEDANCES, 1 OHM.

MODE	ANGULAR OFFSET	130°	140°	150°
TE_{011}	C_o	3.4151×10^{-15}	3.4151×10^{-15}	3.415×10^{-15}
	R_o	0.1657	0.1657	0.1657
	L_o	6.0×10^{-9}	6.0×10^{-9}	6.0×10^{-9}
TE_{112}	C_1	3.8048×10^{-16}	9.7866×10^{-16}	1.769×10^{-15}
	R_1	4.415	1.716	0.9510
	L_1	4.635×10^{-8}	1.8×10^{-8}	1.0×10^{-8}
	C_1'	1.0×10^{-12}	1.0×10^{-12}	1.0×10^{-12}
	L_1'	2.2×10^{-13}	3.8×10^{-13}	4.4×10^{-13}
	M_1	-1	-1	-1
TE_{211}	C_2	5.1785×10^{-15}	1.1×10^{-14}	2.0×10^{-14}
	R_2	0.3275	0.1545	0.08518
	L_2	5.0×10^{-9}	2.3641×10^{-9}	1.3061×10^{-9}
	C_2'	1.5×10^{-13}	7.0×10^{-13}	8.4×10^{-13}
	L_2'	7.0×10^{-13}	7.0×10^{-13}	6.0×10^{-13}
	M_2	-1	1	1
TE_{311}	C_3	1.4×10^{-14}	1.0×10^{-14}	1.226×10^{-14}
	R_3	0.1488	0.2077	0.1690
	L_3	1.24×10^{-9}	1.7254×10^{-9}	1.4×10^{-9}
	C_3'	3.0×10^{-13}	2.7×10^{-13}	5.1×10^{-13}
	L_3'	1.6×10^{-13}	2.6×10^{-13}	2.0×10^{-12}
	M_3	1	1	-1

the above discussion on nulls, which requires a 180° phase shift. Two coupled series resonant circuits with only one resonator shunted by a small capacitor does, however, have the correct phase response and, in addition, has the unusual amplitude response of the experimental data. A physical explanation relating this capacitance to the cavity resonator is not readily apparent, but it may be due to the electric component of the coupling. The use of this circuit does, however, present the possibility of calculating the response of the cavity by utilizing a lumped constant circuit model.

V. Lumped Constant Circuit Model

Bandpass filters, with coupling only between adjacent resonators, can be conveniently analyzed by using lumped constant series resonant circuits with impedance inverters used to adjust the coupling between the resonant circuits [5]. This model can be extended to cylindrical TE_{011} resonators by the addition of resonant circuits to represent the TE_{211} and TE_{311} modes as shown in Fig. 5. Circuit representations of the TE_{211} and TE_{311} modes are placed in parallel with the circuit representing the TE_{011} mode. The TE_{211} and TE_{311} modes are modeled as two series resonant circuits coupled by an impedance inverter and with one of the resonators shunted by a small capacitor. A unity-coupled ideal transformer was placed in each of these circuits to effect the 180° phase shift, relative to the TE_{011} mode, that occurs as the angular offset between the coupling apertures is varied. The relative coupling between the modes is controlled by the L and C ratios.

The transmission characteristics of the TE_{011}-mode resonator were calculated using this circuit model. Since one of the purposes of the calculation was to show correlation with the experimental data, an additional coupled resonator circuit was added to represent the TE_{112} mode. The TE_{112}-mode circuit is not shown in Fig. 5, but it is identical to the TE_{211} and TE_{311} circuit representations and is also connected in parallel. A resistor was also added in series to each resonator to represent the unloaded Q but is not shown in the circuit diagram. Rough circuit values were determined from experimental and other data. The calculations were performed by a Hewlett-Packard 2100S computer using the Opnode circuit analysis program. The circuit values were then modified to produce a calculated response that more closely represented the experimental data. The calculated transmission characteristics, shown in Fig. 6, were obtained using the element values shown in Table I.

The calculated amplitude characteristics of Fig. 6 correlate very well with the experimental data of Fig. 3. The calculated phase characteristics, the dotted line in Fig. 6, clearly show the 180° phase reversals that are necessary to provide the transmission nulls.

VI. Discussion and Conclusions

A powerful method of modifying the frequency response of a single, cylindrical TE_{011}-mode, sidewall-coupled resonator has been presented. Increased selectivity is achieved by creating nulls in the transfer characteristics of the cavity. The existence of a null above or below the TE_{011} response steepens the rejection slope on that side. The nulls are created by making use of the TE_{211} and TE_{311} modes which are naturally excited in the cavity resonator along with the TE_{011} mode. Varying the frequency at which the nulls occur requires control of both the relative phase and amplitude of the modes. The relative phase of the modes is determined by the selection of the angular offset of the coupling apertures, while the relative amplitude of the modes is set by both the angular offset of the apertures and the shaping of the cavity.

A lumped constant equivalent circuit has been presented which is shown to accurately represent the response of the resonator. The equivalent circuit representation can be used as an aid in the design of multisection filters.

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Quarter-Wavelength Coupled Variable Bandstop and Bandpass Filters Using Varactor Diodes

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Abstract —**A quarter-wavelength coupled bandstop filter using varactor diodes for the 6-GHz band has been proposed and tested. Frequency giving maximum attenuation varies from 4.4 GHz–7 GHz. A quarter-wavelength coupled variable bandpass filter using varactor diodes for the 4-GHz band is also proposed and tested. The passband width varies from 730 MHz–1.22 GHz. The center frequency of the filter can also be changed.**

I. Introduction

Many works on the bandstop and the bandpass filters using microstrip have been reported. However, the frequency giving the maximum attenuation in the bandstop filter and passband width in the bandpass filter mentioned above are fixed and cannot be varied.

In this paper, the author proposes a new variable quarter-wavelength coupled bandstop and a variable bandpass filters using varactor diodes. The frequency giving the maximum attenuation of the bandstop filter and passband width of the bandpass filter can be varied mechanically or electrically. These methods of changing the frequency giving the maximum attenuation and passband width have already been reported by the author [1]–[3].

Two types of the filters are considered here. The quarter-wavelength coupled variable bandstop filter is constructed with two circuits which are composed of a series and a parallel resonant circuit connected in parallel, and are placed a quarter-wavelength apart. The frequency giving the maximum attenuation is varied by changing the junction capacitances of varactor diodes mounted in those circuits.

The new variable bandpass filter is composed of two quarter-wavelength coupled bandpass filters connected with coaxial power dividers. Each quarter-wavelength coupled bandpass filter is constructed with two parallel resonant circuits, each of which is composed of a short-circuited transmission line connected with a varactor diode, and is placed a quarter-wavelength apart. The passband width is varied by changing the junction capacitance of those varactor diodes on each quarter-wavelength coupled filter. The experiments were carried out at the 6-GHz band. For the bandpass filter, the passband width was varied from 880 MHz–1.44 GHz. This filter may be used for the tuning reception of the respective signals of the channels of the broadcasting satellite, or for the detection of radar frequencies, and so on. For the bandstop filter, the frequency giving the maximum attenuation was varied from 4.4 GHz–7 GHz. This filter also finds application in broad-band receiving systems which must operate near high power radar, etc.

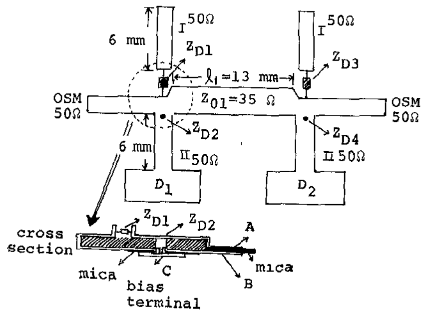

Fig. 1. Quarter-wavelength coupled bandstop filter composed of two composite series and parallel resonant circuits having varactor diodes.

II. Quarter-Wavelength Coupled Variable Bandstop Filter Using Varactor Diodes

The structure of the quarter-wavelength coupled bandstop filter is shown in Fig. 1. A series resonant circuit is structured with a series connection of a short-circuited transmission line I and a varactor diode Z_{D1}, and the parallel resonant circuit is structured with a parallel connection of a short-circuited transmission line II and a varactor diode Z_{D2}. A parallel connection, these two resonant circuits yield a composite series and parallel resonant circuit. Two of the composite resonant circuits are placed a quarter-wavelength apart.

Decreasing in Q-value of the series resonant circuit due to the loading of the varactor diode is recovered by connecting the parallel resonant circuit thereto as shown in Fig. 1, and thereby a narrow bandwidth can be realized.

The bias voltage for Z_{D1} is supplied through terminals A and B, and for Z_{D2} through terminals A and C. The same bias supply

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