Oversized rectangular waveguide components for millimetre waves

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Since microwaves first came into practical use, the hollow pipe, usually of rectangular cross-section, has been the standard means for transmission of the waves. In the conventional use of this waveguide the waves can travel in a single precisely defined pattern — a very valuable property of the guide for many applications. If, however, the same techniques are used in the millimetre and submillimetre wavebands, power losses in the guide becomes a very severe problem and the components become prohibitively small. These have been two of the chief reasons for the development of other types of transmission lines.

In the "oversized waveguide" the losses are much lower than in the standard guide, while the advantage of propagation in a well-defined pattern can be maintained by proper "mode control". The size of the guide can be chosen in a convenient range. Another important advantage is that very broad-band components can be made.

Introduction

At microwave frequencies electromagnetic waves can propagate along several types of transmission line. These can be divided into two classes. In the first class the electromagnetic field extends to infinity in the direction transverse to the axis of propagation; the corresponding structures are usually said to be "open". In the second class the electromagnetic field has only a finite transverse extension. Since the boundaries are imposed by metallic screening, this class is said to be "shielded" or "closed".

Well-known transmission lines such as the dielectric line [1], the image line [2], the H-guide [3] and the microwave beam [4], belong to the first class. The shielded structures of the second class include hollow pipes and pipes containing one or more inner conductors. The rectangular waveguide and the coaxial line are typical representatives of this class.

In the millimetre-wave region, to which we shall confine ourselves, both classes of structures have their specific fields of application and both have their special advantages [5]. In this article we shall be concerned with the closed structures. Let us begin by noting some of their advantages. First, the shielding prevents radiation and makes the device insensitive to parasitic electromagnetic fields; there is no coupling even between closely spaced guides. Secondly, the shielding part of the structure gives sufficient rigidity without requiring additional support, and alignment is quite easy. Finally, many almost ideal components (bends, directional couplers, etc.) can be produced: this is not so with the open structures, where diffraction phenomena may easily give rise to unpredictable deviations from the ideal behaviour.

A traditional closed microwave structure is designed to operate in a single mode: waves of a single definite pattern only can be propagated. For this to be true the structure must be of "standard size", i.e. the cross-sectional dimensions must not exceed certain critical values which are of the order of magnitude of the wavelength used. For millimetre waves in particular, if a cross-section is to be of standard size it must be smaller than a few millimetres across. Coaxial transmission lines of this size are extremely lossy and therefore not very suitable for millimetre-wave applications. On the other hand, standard-size rectangular waveguide components for wavelengths as short as 2 mm have been developed and are in current use [6].

If, at a given frequency, the cross-sectional dimensions of a guide do exceed the critical value referred to above, transmission is possible in several modes. As we shall explain more fully below, this will in general, lead to an unstable electromagnetic field pattern, that is to say, the pattern may vary erratically along the guide. If, however, it is possible to prevent the excitation of the higher-order modes, such "oversized" transmission lines have several advantages over standard-size lines. To mention a few: the attenuation can be considerably lower, broadband components can be

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designed, and, particularly in the millimetre and sub-
millimetre range, manufacture can be easier. The spec-
fic problems associated with oversized transmission
lines will be the subject of the present article.

Transmission lines of three main types of cross-
section are usually considered. These are: the rectangu-
lar waveguide, the circular waveguide, and the coaxial
line. The latter is of no interest here, as the design of
components not exciting higher-order modes proves to
be very difficult. The oversized circular waveguide has
become very promising, particularly for long-distance
transmission [9]. For laboratory applications, however,
the rectangular form offers certain advantages.

The present article is mainly about the rectangular
type of oversized waveguide, which has already attracted
some interest [8]; a brief comparison between the cir-
cular type and the rectangular type will be given at the
end. After a general discussion of the mode spectrum
of the rectangular waveguide, we pay special attention to
the measurement and control of higher-order modes.
Several components which have been developed in
our laboratory will be described in detail. In particular,
we discuss mode filters, tapers, bends, directional
couplers, and variable attenuators, phase shifters, and
impedances. Finally various applications are enu-
merated.

Most of our components have been designed for and
tested in the 4-mm wave region, and have been built
using standard 3-cm waveguide as a starting point.

Mode spectrum of the rectangular waveguide

To give an insight into the problems associated with
higher-order modes let us consider some fundamental
properties of hollow waveguides [9]. Assuming that we
have a waveguide of constant cross-section, let us raise
the frequency (instead of taking increased dimensions at
a given frequency, as above). To begin with no propa-
gation of electromagnetic waves is possible at all below
a critical frequency, the cut-off frequency $f_{c1}$. This cut-
of frequency is a specific property of the cross-section
under consideration. For frequencies $f$ slightly above
$f_{c1}$ the electromagnetic field propagates in a charac-
teristic pattern, the fundamental mode of the waveguide.
If the frequency is raised further, a second cut-off fre-
quency $f_{c2}$ is obtained, above which another mode with
a different field pattern can be propagated. (For $f_{c2}$
the waveguide has the critical dimensions referred to in
the introduction.) Every waveguide has an infinite
spectrum of such cut-off frequencies $f_{c3}, f_{c4}, f_{c5}, \ldots$, so that for sufficiently high frequencies the energy can
be transmitted in a large number of modes simultane-
ously. In most cases it turns out that the cut-off wave-
length $\lambda_{c1}$ corresponding to the first cut-off fre-
quency $f_{c1}$ has the same order of magnitude as the di-

ections of the cross-section (this wavelength is given by
$\lambda_{c1} = c/f_{c1}$, where $c$ is the free-space velocity of
light). Thus any wave that can be propagated has a
free-space wavelength $\lambda$ comparable to these dimen-
sions or smaller.

A specific mode is propagated along the waveguide
without change of the transverse field distribution. In a
lossless waveguide, the amplitude is constant. The phase
velocity $v_i$ of the mode $i$, i.e. the velocity of propaga-
tion of the field pattern, is given by:

$$v_i = \frac{c}{\sqrt{1 - (\lambda/\lambda_{ci})^2}}, \ldots \ldots (1)$$

where $\lambda_{ci}$ is the cut-off wavelength of the mode. Thus,
at a given frequency, the phase velocities are different
for different modes. Therefore, if several modes are
excited, the transverse electromagnetic field distribution
varies along the guide owing to the continuously vary-
ing phase differences between the individual modes.
This unstable situation may occur if $f > f_{c2}$, and if
the frequency satisfies this condition the waveguide is
said to be overmoded or oversized. If the frequency lies
between $f_{c1}$ and $f_{c2}$, the case which we shall refer to as
standard-size operation, only the fundamental mode is
present and the propagation is stable. If stable opera-
tion is required in oversized waveguide, special care
has to be taken to excite only one mode, which need not
necessarily be the fundamental one. Such "mode control" will be discussed in detail in the next section.

We now apply these general considerations to the
special case of a rectangular cross-section. Let $a$ and

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sion lines for microwave frequencies, Philips tech. Rev. 26,

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article in ref. [1].

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378-394, 1964 and Nachr.techn. Z., edn. Comm. J. 6, 264-
269, 1965.

[5] An extensive survey of several types of transmission line
studied in view of their application at millimetre and sub-
millimetre wavelengths and paying special attention to
losses has been made by D. J. Kroon and J. M. van Nieuw-
land of this laboratory. This survey is published in: D. H.
Martin, Spectroscopic techniques for far infra-red, submilli-
metre and millimetre waves, North-Holland Pub!. Co.,
Amsterdam 1967, Chap. 7: Techniques of propagation at
millimetre and submillimetre wavelengths, p. 308-380.

equipment for 2 mm microwaves, Philips tech. Rev. 22,


[9] For extensive theoretical treatments of this subject the reader
is referred to standard textbooks on waveguide theory. An
introduction is given in: W. Opechowski, Electromagnetic
waves in waveguides, Philips tech. Rev. 10, 13-25 and 46-54,
1948/49.
Fig. 2. Field pattern of the fundamental mode and some higher-order modes in rectangular waveguide. The longitudinal cross-section is shown above, and the transverse cross-section below. The electric field lines are shown solid and the magnetic field lines dashed. It should be noted that many of the lines drawn in the cross-sectional planes are not the actual field lines, but projections of the field lines on those planes. In the longitudinal cross-sections the directions indicated (arrows) are those of the lines near the top of the guide, and in the transverse cross-section the directions are those of the lines near the section A-A (as indicated in the first drawing). These directions correspond to the direction of propagation indicated between the first two drawings. The various modes are drawn in such a way that they have the same guide wavelength \( \lambda_g = \frac{v_I}{f} \) (the length of the longitudinal section drawn is ~\( \lambda_g \)); this implies that, if the \( \lambda_{cl} \)'s differ, the frequencies will be different (cf. eq. 1).

\[ b \] be the width and the height of the rectangle, respectively, where \( a > b \). The first cut-off wavelength is then given by:

\[ \lambda_{c1} = 2a. \]  \hspace{1cm} (2)

The field pattern of the fundamental mode, often referred to as the \( H_{10} \) mode, is shown in fig. 1. The electric field lines \( E \) are parallel to the shorter side of the waveguide while the magnetic field lines \( H \) lie in planes perpendicular to the electric field. The cut-off wavelengths \( \lambda_{cnm} \) of the higher-order modes are given by:

\[ \frac{2}{\lambda_{cnm}}^2 = \frac{(m/a)^2}{(n/b)^2}. \]  \hspace{1cm} (3)

This expression yields the complete mode spectrum: an infinite set of modes, each characterized by two mode numbers \( m \) and \( n \), one of which, but not both, may be zero.

In fig. 2 the field patterns of the fundamental mode and some higher-order modes of the rectangular waveguide are shown. If \( m = 0 \) or \( n = 0 \), then only the magnetic field has an axial component \( H_z \); the axial component of the electric field \( E_z \) vanishes. Modes of this type are generally called \( H \) modes. Thus we have an \( H_{90} \) mode, an \( H_{90} \) mode, an \( H_{90} \) mode, etc. If neither \( m \) nor \( n \) is zero, two different field patterns are possible which have the same cut-off wavelength \( \lambda_{cnm} \). The first, known as \( H_{mn} \), has again a vanishing electric field component \( E_z \), but a non-vanishing \( H_z \); the second, known as \( E_{mn} \), has a vanishing magnetic field component \( H_z \), but a non-vanishing \( E_z \). The indices \( m \) and \( n \) are equal to the number of half periods of the wave pattern along the width \( a \) and the height \( b \) respectively.

Fig. 3 is an illustration of how the field pattern of the fundamental mode in an oversized waveguide is elongated in the transverse direction and compressed in the longitudinal direction, compared with the pattern in standard-size guide. When a guide is highly oversized the pattern is close to that of a transverse electromagnetic wave.

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Eq. (3) yields the cut-off frequencies \( f_{c_mn} = c/\lambda_{c_mn} \) for any combination of indices \( m \) and \( n \). The mode spectrum \( f_{c1}, f_{c2}, f_{c3}, \ldots \), arranged in order of increasing cut-off frequencies, depends upon \( a \) and \( b \). The ratio \( a/b \) determines the higher cut-off frequencies relative to \( f_{c1} \). Mode spectra of this type are shown in fig. 4 for the two values \( a/b = 22.86/10.16 = 2.25 \) and \( a/b = 22.86/1.55 = 14.8 \). The first case corresponds to standard 3-cm waveguide (inner dimensions \( 0.9'' \times 0.4'' \), i.e. \( 22.86 \times 10.16 \text{mm} \)), the second to a flat waveguide, with the same width as 3-cm waveguide but height equal to that of 4-mm waveguide. For both configurations the ratio \( f_{c2}/f_{c1} \), which determines the frequency band of stable \( H_{10} \) mode transmission, is equal to 2. It is easy to show that this is the case whenever \( a/b \geq 2 \). If this condition is met the rectangular waveguide, for standard-size operation, can be used theoretically over a frequency range of an octave. For practical purposes, however, only part of this band is useful: in the vicinity of \( f_{c1} \) the losses of the \( H_{10} \) mode become very high, and in the vicinity of \( f_{c2} \) the next higher-order mode \( H_{20} \) is almost propagating, which means that this mode, if excited by an obstacle, dies away very slowly. The frequency ratio \( f_{max}/f_{min} \) of the

\[ m_0 \quad m_1 \quad m_2 \quad m_3 \quad m_4 \]

\[ \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \]

\[ 3n \quad 2n \quad n \]

\[ \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \]

Fig. 4. Mode spectrum of standard 3-cm waveguide (upper diagram; \( a/b = 2.25 \)) and spectrum of a flat waveguide with the same width as 3-cm waveguide but height equal to that of 4-mm waveguide (lower diagram; \( a/b = 14.8 \)). The spectra show the cut-off frequencies \( f_{c_mn} \) above which the mode \((m, n)\) can propagate. The numbers \( m \) and \( n \) are indicated in the figure. For a specific pair \( m, n \), both \( \neq 0 \), two modes exist: the \( E_{mn} \) mode and the \( H_{mn} \) mode, with the same cut-off frequency. For \( m \) or \( n = 0 \) only the modes \( H_{0n} \) and \( H_{m0} \) are possible; \( E_{0n} \), \( E_{m0} \) and \( H_{00} \) do not exist.
band recommended for standard-size operation of rectangular waveguides is only 1.6.

Mode control

As mentioned above, special care has to be taken if, in an oversized waveguide, a single mode is wanted. This is because, even when a single mode is launched at the input, many modes will easily arise in a circuit: in each microwave component and even at each irregularity in a waveguide there will be mode conversion, i.e. if a wave of a certain mode is incident on the component or at the irregularity, other modes are excited.

One may ask just why multimode operation is so undesirable. The answer is that in multimode operation a simple description of the electromagnetic waves in a circuit is no longer possible. Even in a simple piece of oversized waveguide the transverse field distribution is not well defined and varies along the guide. More generally, if a component is considered as a black box with a number of waveguide arms, the reflection and transmission coefficients of the wanted modes as well as the conversion coefficients must be known for a complete description of such a component. If there are several components in the circuit, the mathematical description may soon become so unwieldy that in practice the field distribution is unpredictable; on the other hand, measurements in such a circuit will not lead to definite conclusions about the characteristics of a certain component.

If the conversion from the wanted mode to unwanted modes is negligible, we can refer to single-mode or stable operation of a specific oversized waveguide circuit. However, it should be emphasized that, even if the coefficients describing the conversion into the separate components are very small, higher-order modes with substantial amplitudes may arise. This may occur for example if the higher-order modes, excited at some irregularity, are trapped between two tapers, as shown in fig. 5. A taper (to be treated in the following section) gives a smooth transition between an oversized waveguide and a standard-size waveguide. Since only the fundamental mode can propagate in the standard guide, a higher-order mode incident from the oversized section is totally reflected. Multiple reflections of the higher-order modes may give rise to resonance of these modes. If the frequency of the source coincides with one of the resonance frequencies, absorption takes place causing a considerable loss in the total transmitted power. This, when plotted as a function of frequency (fig. 5), may exhibit a great number of such resonance absorptions. The density of these resonances increases with increasing degree of oversizing and increasing length of the trapped-mode resonator. Furthermore, the larger the conversion coefficient of the irregularity and the lower the attenuation per unit length of the resonating mode, the stronger the absorption. In the limiting case of vanishing attenuation the transmission may even be zero [10], which corresponds to a total reflection of the incident wave.

There are some special applications of oversized waveguides where it would be very difficult to achieve single-mode transmission, but multimode operation is however acceptable. For instance, if submillimetre waves are transmitted through a standard centimetre-waveguide, this guide is oversized to a very high degree, and in such a case it is extremely difficult to make an almost ideal taper and launch only a single mode [11]. Nevertheless, such a guide may be useful, if only for transferring power over normal laboratory distances with low attenuation. In this case the existence of higher-order modes does not essentially affect the overall transmission, for if the waveguide is long enough the trapped-mode resonances are well damped out.

Such special cases apart, oversized waveguides will usually be used in single-mode operation. This is almost invariably so in measurement work. If, for instance, a bridge circuit has to be designed, the components, such as power dividers and directional couplers should be substantially free from mode conversion, since the excitation of higher-order modes may easily lead to unpredictable errors.

![Fig. 5. Trapped-mode resonances. Microwave power from a swept-frequency generator 1 with standard-size output is fed through a taper 2 into oversized waveguide 3 and by a second taper 4 into a broadband detector 5 in standard-size waveguide. The tapers are assumed to be free from mode conversion. Mode conversion will take place at any irregularity 6. A higher-order mode will be reflected in the tapers and be trapped, giving resonance losses at certain frequencies, which can be determined from the power-frequency plot on the oscilloscope 7. a) Low mode conversion at 6; b) high mode conversion; c) the density of resonances increases if the length-of the oversized waveguide 3 or its degree of oversizing is increased.](image)
Some time ago, we began an investigation in this laboratory to find out how to design components with low mode conversion. For this work, we needed equipment for measuring the conversion coefficients. An essential element of such equipment is the mode transducer, several types of which are described below. If the components to be developed do not satisfy certain requirements for mode purity, this can be improved by mode filters, which will be described later. The measurements to test the components that had been developed were all made with the aid of a swept-frequency technique in which we used a backward-wave oscillator that could be swept between 72 and 77 GHz.

Mode transducers

A typical arrangement for the measurement of mode conversion using a mode transducer is shown in fig. 6. The waveguide component under test is excited by a wave in the wanted $H_{10}$ mode incident from the left. An $H_{10}$ mode and several higher-order modes produced in the component by mode conversion leave by the right-hand arm. All these modes except the one to be measured ($X$) are completely absorbed by a mode filter before the transducer. The mode filter is often incorporated in the transducer. The remaining mode $X$ is converted into the $H_{10}$ mode in the outgoing standard-size waveguide of the transducer. The way in which this conversion is actually performed is different for each type of mode transducer, and will be discussed below. If the transmission characteristics of the mode transducer are known the overall $H_{10} \rightarrow H_{10}$-transmission of the arrangement is a measure of the mode conversion $H_{10} \rightarrow X$ in the component under test. Care must be taken that the transducer does not reflect the mode $X$ to be measured, since a reflected wave might cause further mode conversion in the unknown device. The input waveguide of the component under test may be oversize or of standard size.

In the component method proposed by Peau de Cerf and by Levinson and Rubinstein. This method is based on probing the electromagnetic field with the aid of a single movable probe or a number of identical fixed probes. The amplitudes of the individual modes can be evaluated from the field distribution measured in this way. The probe method has the advantage that the mode distribution can be measured in a system while it is operating, while the method using mode transducers only permits the evaluation of the mode conversion in a single component. Small distortions of the wanted mode pattern, however, cannot be accurately determined by the probe method, whereas with the transducers to be described below, spurious mode levels of more than 40 dB below the fundamental mode level can easily be measured. Another disadvantage of the probe method is that the probes may disturb the original field distribution.

Before describing a few types of mode transducer in detail, let us consider the possible higher-order modes in a rectangular waveguide. These modes (cf. fig. 2) can be divided into two classes according to whether the field does or does not vary in the y-direction (the direction perpendicular to the broad face of the guide).

Class 1, in which the field does vary with $y$, consists of the $E_{mn}$ and $H_{mn}$ modes with $n \neq 0$. Class 2, in which the field does not depend on $y$, contains the $H_{mn}$ modes with $m \neq 1$. It will be shown that mode filters can easily be designed for the modes belonging to class 1. Thus no difficulties arise from conversion into these modes provided the power loss of the $H_{10}$ mode is not excessive. For this reason, mode transducers have been designed for the second class only, and in this class only for the most interfering types $H_{20}$ and $H_{10}$.

An $H_{10} \rightarrow H_{50}$ mode transducer first proposed by Montgomery, Dicke and Purcell, is shown in fig. 7. A standard-size waveguide is split into two parts by means of a metal sheet perpendicular to the electric field. The two guides then twist through 90° in opposite senses. This brings the two field patterns side by side, giving the wanted $H_{20}$ mode. Finally there is a smooth linear taper to an oversized waveguide. As this device is reciprocal it can be used in both directions, i.e. as an $H_{20}$ detector or as an $H_{20}$ exciter. This transducer is subject to some mode conversion in the taper, but this effect can be made arbitrarily small by using a suffi-

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ciently smooth taper. Furthermore, a small asymmetry in the twisted section may give rise to a residual $H_{10}$ mode in the oversized waveguide; this is absorbed by means of a resistive sheet placed at the centre of the guide. An additional function of this sheet is the absorption of any $H_{10}$ mode energy that might arrive from the oversized section. Finally, care must be taken to avoid trapped-mode resonances associated with modes which cannot propagate through the taper. It is therefore advisable to use this transducer in connection with one of the separate mode filters described in the following section.

We have developed a mode transducer for the $H_{30}$ mode, making use of the principle of mode-selective directional coupling (fig. 8). Two waveguides whose widths differ by a factor of 3 are coupled by means of a slot. Since the $H_{10}$ mode in the narrow guide and the $H_{30}$ mode in the broad guide have equal phase velocities they interact strongly. In a way analogous to the action of two coupled pendulums of equal resonance frequency, the energy of the $H_{10}$ mode in the narrow guide is transferred into the $H_{30}$ mode in the broad guide, the distance for complete transfer depending on the width of the slot. A smooth taper guides the $H_{30}$ mode into the oversized waveguide.

Fig. 8. Transverse cross-section of $H_{10} \leftrightarrow H_{30}$ mode transducer. Owing to the equal phase velocities of the $H_{10}$ mode in the upper guide and the $H_{30}$ mode in the lower guide, power in one of these modes will be completely transferred into the other in a distance along the guides which depends on the width of the coupling slot.

Measurements on a complete mode-transducer assembly, using a 12-μm "Mylar" sheet with a resistive metallic film for the mode filter, have indicated an $H_{30}$ mode purity of better than 50 dB. The surface impedance of the resistive sheets of the mode filter can be optimized by applying well-known principles described elsewhere [13]. For standard X-band (3-cm) waveguide...
and a frequency $f = 75$ GHz (4 mm) one obtains an optimum surface resistance of $R = 560$ ohms giving an attenuation of $a = 2.4$ dB/cm for the $H_{10}$ mode. As in the $H_{90}$ mode filter the resistive sheets have the additional function of absorbers for any $H_{10}$ mode energy incident from the oversized waveguide section. These sheets also suppress potential trapped-mode resonances.

**Mode filters**

In the mode transducers described above suitable mode filters were incorporated where necessary. As, however, mode conversion is an effect which can easily occur in oversized components, separate mode filters are desirable for mode control. A mode filter is essentially a two-port which strongly attenuates unwanted modes, while transmitting the wanted mode (usually the $H_{10}$ mode) almost unperturbed. Furthermore, all modes should be matched; in other words, the mode-selective transmission should not be achieved by means of reflecting obstacles as these might give rise to trapped-mode resonances.

A simple design for a mode filter for the higher-order modes of class 1 (cf. p. 91), that is to say, the $E_{mn}$ and $H_{mn}$ modes with $n \neq 0$, consists of one or more resistive sheets perpendicular to the electric field lines of the wanted $H_{10}$ mode (fig. 9). Such sheets do not perturb the propagation of the $H_{10}$ mode, but attenuate $E_{mn}$ and $H_{mn}$ modes with $n \neq 0$ since these modes have an electric field component parallel to the sheets. Unperturbed transmission can however occur if the planes of the sheets happen to coincide with nodal planes of one of these modes.

It should be noted that an $H_{mn}$ mode ($n \neq 0$) incident at one port of this general type of mode filter gives not only an (attenuated) $H_{mn}$ mode at the other port, but an $E_{mn}$ mode as well (with the same subscripts $m$, $n$). Likewise, an incident $E_{mn}$ mode sets up an $H_{mn}$ mode at the other port. This mode conversion is related to what is called the "degeneracy" of the two modes $E_{mn}$ and $H_{mn}$, which means that their phase velocities are equal. If the modes $E_{mn}$ and $H_{mn}$ are superimposed in a rectangular waveguide, their phase difference does not vary along the guide, i.e. the field pattern is stable. Thus every superposition of these modes also gives a mode.

It has been shown \[15\] that there are two such combinations of the $E_{mn}$ and $H_{mn}$ mode which undergo no mode conversion in this type of mode filter. These combinations are characterized by the properties $E_y = 0$ and $H_y = 0$, respectively. This implies that either the electric or the magnetic field lines lie in planes parallel to the sheet. These two modified $mn$-modes are often called "longitudinal-section modes" (LS-modes). An $E_{mn}$ or $H_{mn}$ wave incident from the empty guide sets up both types of LS-mode. Thus for an efficient filtering of $E_{mn}$ and $H_{mn}$ modes both LS-modes have to be attenuated sufficiently in the mode filter.

The electric field patterns of the LS-modes inside the mode filter are shown in fig. 10 for the case $m = n = 1$ and with one symmetrically located sheet. The

\[\text{\[15\]}\] H. J. Butterweck, Mode filters for oversized rectangular waveguide, to be published in IEEE Trans. MTT.

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first mode, called \( EH_{11} \), has two electric field components, \( E_x \) and \( E_z \), parallel to the sheet, \( E_z \) being negligible for highly oversized waveguide. This mode is considerably attenuated by the sheet. For the second mode, \( EH_{11} \), however, the only important component parallel to the sheet is \( E_z \) and this axial component is very small anyway, as might be expected since waves in highly oversized waveguide are approximately transverse electromagnetic in character (cf. fig. 3). Thus the surface resistance of the sheet has to be chosen carefully to obtain a reasonable value for the attenuation of the \( EH_{11} \) mode. As is usual in absorption problems, there is an optimum value of the surface resistance \( R_\square \) (neither a perfect conductor nor a perfect insulator absorb electromagnetic waves), and for the case of a single symmetrically located sheet this value has been shown to be \( 15 \): 
\[
R_\square = 140 \lambda/b \text{ ohms.} \quad (4)
\]

The corresponding maximum attenuation of the \( EH_{11} \) mode (in dB per unit length) is:
\[
\alpha_{\text{max}} = 3.8 \lambda/b^5. \quad (5)
\]

For our example of standard 3-cm waveguide \((b = 10.16 \text{ mm})\) and \( \lambda = 4 \text{ mm} \), we obtain \( R_\square = 56 \text{ ohms} \) and \( \alpha_{\text{max}} = 1.5 \text{ dB/cm} \).

It should be mentioned that the comparatively low value of \( R_\square \) given by eq. (4) is quite different from the corresponding optimum value for the \( EH_{11} \) mode. For a high degree of oversizing, the attenuation of the \( EH_{11} \) mode can be even lower than that of the \( EH_{11} \) mode, if the optimum value for the latter is chosen from eq. (4). In this case a compromise for the value of the surface resistance \( R_\square \) is recommended \(15\).

For the higher-order modes of class 2 (p. 91), that is to say the \( H_m \) modes with \( m \neq 1 \), we were unable to design a mode filter with such an almost ideal performance as that of the filter described above. However, we give details of two designs which do at least meet certain requirements. The first type is inherently broadband and attenuates all the higher modes simultaneously; but the wanted \( H_{10} \) mode is also attenuated to a certain extent. The second type of mode filter transmits the \( H_{10} \) mode without losses, but is narrow-band and absorbs only a specific higher-order mode.

The first design consists of a waveguide whose narrow walls are made of lossy material (see fig. 11). If the broad walls are assumed to be perfectly conducting, the attenuation (in dB per unit length) of the \( H_m \) modes is given by
\[
\alpha_m = 8.67 \frac{m^2 \lambda^2}{2a^3} \frac{v_m}{c} \frac{R_w}{Z_0}. \quad (6)
\]

Here \( v_m \) is the phase velocity of the mode under consideration as given by eq. (1), \( R_w \) is the surface resistance of the lossy walls, and \( Z_0 = 377 \) ohms is the free-space characteristic impedance. If the waveguide is highly oversized and if \( m \) is not too high, we have \( v_m \approx c \) so that \( \alpha_m \) may be assumed to increase in proportion to \( m^2 \). Thus the \( H_{10} \) mode is attenuated nine times as much as the \( H_{10} \) mode.

This dependence of the attenuation upon the mode number \( m \) can be illustrated with the aid of the well-known representation of an \( H_m \) mode by the zigzag reflections of a plane wave \( 14 \).

The angle \( \Theta \) (see fig. 11) between its direction of propagation and the reflecting wall follows from the relation:
\[
\sin \Theta = \frac{\lambda}{2m} = \frac{m \lambda}{2a}. \quad \quad \quad (7)
\]

Thus \( \Theta \) increases with increasing mode number \( m \). The distance \( L \) which the wave travels along the guide between two reflections is given by:
\[
L = a \tan \Theta. \quad \quad \quad (8)
\]

The relative power loss per reflection equals \( 1 - |r|^2 \), where \( r \) is the complex amplitude reflection coefficient of the wall. The power loss in dB per unit length of the guide, which equals twice the attenuation constant \( \alpha_m \), is therefore:
\[
2\alpha_m = 8.67(1 - |r|^2)/L = 8.67(\tan \Theta)(1 - |r|^2)/a. \quad (9)
\]

The power reflection coefficient \( |r|^2 \) of a plane wave obliquely incident on a dissipative medium with surface resistance \( R_w \) can be found in standard textbooks and is given by:
\[
|r|^2 = 1 - 4(R_w/Z_0)\sin \Theta. \quad \quad \quad (10)
\]

If eqs. (7) and (10) are substituted in (9) and use is made of eq. (1), we obtain eq. (6) for the attenuation of the \( H_m \) modes. It is clear that the \( m^2 \) law is due to the fact that a) the number of reflections per unit length, and b) the power loss per reflection increase with increasing mode number \( m \).

\[\text{Fig. 11. Filter for the modes } H_m \text{ with } m \geq 2. \text{ The side walls are made of lossy material. This filter is not ideal in principle as the } H_{10} \text{ mode undergoes some attenuation, but the attenuation increases with the mode number } m \text{ (cf. eq. 6). To illustrate the derivation of eq. (6) an } H_0 \text{ mode is represented by the plane wave } I \text{ which is reflected in a zigzag path by the narrow walls.}\]

The second type of mode filter, designed for the suppression of a specific \( H_m \) mode in a highly oversized waveguide, is shown in fig. 12. The filtering action is based on the principle of mode-selective directional coupling. Two subsidiary waveguides are coupled to
the main guide by means of dielectric sheets. Let the mode to be absorbed by the filter be of the \( H_{30} \) type. In this case the width of the subsidiary guides is made about a third of that of the main guide. The phase velocity of the \( H_{10} \) mode in these subsidiary waveguides then approximately equals the phase velocity of the \( H_{30} \) mode in the main guide, resulting in a strong interaction between these two modes. As in the \( H_{10} \rightarrow H_{30} \) transducer described in the preceding section, all the power of the \( H_{30} \) mode in the main guide can be transferred to the subsidiary guides and dissipated there, provided all the parameters (physical dimensions, dielectric constant of the sheet) are related in a certain manner which we shall not go into here. A detailed theoretical treatment is given elsewhere [15].

![Diagram of filter for the \( H_{30} \) mode.](image)

Both types of \( H_{m0} \) mode filter described above suffer from the general disadvantage of mode conversion from the \( H_{10} \) mode to higher-order modes \( H_{m0} \) with \( m \) odd. This effect is also described in reference [16].

Mode filters of the three designs described have been made in standard 3-cm waveguide and their performance in the 4-mm wave region has been measured. In the first type (cf. fig. 9) we used three equidistant resistive sheets with a surface resistance of 70 ohms and a length of 10 cm. We are not able to quote quantitative results since we had no suitable mode transducer, but the complete suppression of strong trapped-mode resonances did give a rough indication of an effective mode filtering. The second type (fig. 11) was constructed with a lossy wall having a surface resistance of about 250 ohms. With a length of 15 cm we obtained the following attenuations: 0.6 dB, 2.5 dB, and 5.4 dB for the \( H_{10} \), \( H_{30} \), and \( H_{30} \) modes, respectively. A filter of the third type (fig. 12), with a length of 15 cm, gave attenuations of 0.3 dB and 30 dB for the \( H_{10} \) and \( H_{30} \) modes respectively. For both of the last two types of filter the mode conversion \( H_{10} \rightarrow H_{30} \) was about \(-30 \) dB.

**Components**

We shall now describe several components for oversized rectangular waveguide, some of them new, paying regard to mode conversion. Special attention will be paid to their broadband characteristics. In most cases we were able to design components for a 10 to 1 frequency band with the lowest frequency corresponding to standard-size operation.

**Tapers**

The waveguide output of coherent microwave sources is usually designed for standard-size operation. This means that some sort of transition is needed from standard-size to oversized waveguide. In order to launch an almost pure \( H_{10} \) mode in the oversized waveguide, this transition has to be very smooth. Such a "taper" is also needed for the normal types of detector mounted in standard-size waveguide. In principle, a detector can also be designed for oversized waveguide, e.g. in the form of a homogeneous detecting sheet mounted transversely in the waveguide. However, such a detector is too insensitive for most applications. For good sensitivity some sort of concentration of the energy is required. This can be achieved with a taper much more efficiently than with other devices such as concave mirrors and lenses, which moreover will excite higher modes in the oversized guide.

In a taper the field pattern of the \( H_{10} \) mode is elongated in the transverse direction and compressed in the longitudinal direction. This must not involve a substantial perturbation of the characteristic field pattern of the mode; in other words, no serious mode conversion is allowed to occur. Furthermore, the reflection coefficient of a taper must be as low as possible. Fortunately in most cases, tapers optimized for low mode conversion also automatically exhibit negligible reflection coefficients. Measurements have indicated that the reflection coefficient is of the order of 1%.

The tapers considered here have a smoothly-varying cross-section with a) every cross-section rectangular, b) the centres of all the rectangles on a straight line, the "axis" of the taper, c) no twisting of the rectangles with respect to each other. It has been shown [11] that in a smooth rectangular waveguide taper the excitation of the \( H_{30} \) mode, which is found to be the most troublesome one, is dependent only upon the axial variation of the width \( a \) of the taper. Roughly speaking, the mode

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conversion decreases with increasing length of the taper, but for a given length the form of the taper can be optimized in order to yield minimum mode conversion.

In the theoretical design providing a minimum $H_{10} \rightarrow H_{30}$ mode conversion the function $a(z)$ has a bottle-like shape as shown in fig. 13; $b$ can be a linear function of $z$. A taper of this form, 140 mm long, between standard 4-mm waveguide (1.55 \times 3.10 mm inner dimensions) and standard 3-cm waveguide, gave a mode conversion $H_{10} \rightarrow H_{30}$ of less than $-40$ dB (i.e. the excited $H_{30}$ mode is at least $40$ dB below the $H_{10}$ mode). On the other hand, a taper with a linear function $a(z)$ and the same length gave a much larger $H_{10} \rightarrow H_{30}$ conversion (a value of about $-18$ dB was measured).

![Fig. 13. The taper for transferring the $H_{10}$ mode from standard-size to oversized waveguide. The design is based on a minimum $H_{10} \rightarrow H_{30}$ mode conversion for a given length of the taper. The width $a$ is a function of $z$ with a bottle-like shape, the height $b$ a linear function of $z$.](image)

We can understand the superiority of the “bottle” form shown in fig. 13 as follows: owing to the finite inclination of the walls an infinitesimal element of the taper at $z = z_0$ gives rise to two elementary waves of the unwanted $H_{30}$ mode; one travels towards the oversized section (to the right in fig. 13), the other in the opposite direction. The wave travelling to the left is totally reflected in the cross-section at $z = z_c$, for which the $H_{30}$ mode becomes cut-off, then it too travels, as a secondary wave, toward the oversized section. Now, for the primary waves, let us consider the phase difference $\Delta \varphi$ between the $H_{10}$ wave incident at the origin ($z = 0$) and the $H_{30}$ wave leaving at the end ($z = l$). $\Delta \varphi$ is determined by the phase velocity of the $H_{10}$ wave from $z = 0$ up to $z = z_0$ and that of the $H_{30}$ wave from $z = z_0$ up to $z = l$. It is therefore dependent upon the position $z_0$ of the infinitesimal mode-converting element under consideration. Obviously the total $H_{30}$ amplitude found by integration of the elementary contributions is far less than the sum of the absolute values because of “destructive interference” (i.e. contributions of opposite phase cancel each other, partially or totally, depending on their magnitudes). Since the phase constants (phase shifts per unit length) of the $H_{10}$ and $H_{30}$ mode both approach the free-space phase constant with increasing width $a$ of the taper, their difference tends to zero as $a$ tends to infinity. The destructive interference therefore becomes less effective with increasing width, which means that the contributions from the end of the taper, corresponding to the largest values of the width $a$, have a relatively large weighting factor. This part therefore has to be designed with a very small slope to ensure small elementary contributions to the mode conversion, whereas the narrower part may have a greater inclination. This approach leads to the “bottle” shape for the taper. With the secondary waves reflected at the cut-off plane, the sum of their two phase constants determines the destructive interference; this leads, after integration, to total amplitudes which are always negligible when compared with the effect of the primary waves.

An important concept in connection with the analysis of tapers and other oversized waveguide components is the beat wavelength $\lambda_b$ of two distinct modes, i.e. the length along which the phase difference of the two modes changes by $2\pi$. In other words $\lambda_b$ equals $2\pi$ divided by the difference of the two phase constants. To obtain sufficient destructive interference in a component, it should be long with respect to the beat wavelength $\lambda_b$. For two specific modes $\lambda_b$ is dependent on the frequency and the dimensions $a$ and $b$. Thus, for the modes $H_{10}$ and $H_{30}$, $\lambda_b = 130$ mm for $f = 75$ GHz and standard 3-cm waveguide.

Because of the symmetrical structure of this kind of taper $H_{mn}$ modes with $m$ even or $n$ odd cannot be excited, and therefore only the types $H_{30}$, $H_{50}$, ..., $H_{13}$, $H_{35}$, ..., $H_{14}$, $H_{34}$, ... occur. Of these, the $H_{30}$ mode usually has the largest amplitude, since the corresponding beat wavelength $\lambda_b$ is by far the largest. The tendency towards destructive interference is therefore particularly small with this type of mode, and on this account
our taper has only been optimized for low $H_{10} \rightarrow H_{30}$ mode conversion.

The only part of the taper which is significant in mode conversion is the part where the unwanted higher mode can propagate. The rest of the taper can be designed more or less arbitrarily: we chose a linear function for $a(z)$. We made the variation of $b$ with $z$ linear throughout, since variation in height excites only the types $H_{18}, H_{38}, \ldots, H_{14}, H_{34}, \ldots$ which always have small amplitudes on account of their small beat wavelengths.

**Bends**

There are two different approaches to the design of the majority of the components for oversized waveguide. The first, used by most workers in this field \[8,17,171,\] is the "optical approach", in which the propagation of electromagnetic energy is assumed to take place essentially along "rays", which undergo reflection and refraction, as in geometrical optics. This approach is useful especially for a high degree of oversizing, since in this case the fundamental $H_{10}$ wave behaves very much like a plane wave. However, there are some components such as tapers and mode filters which cannot be designed by using optical principles. It has been found very difficult to make these indispensable elements with low enough mode conversion if the waveguide is highly oversized. In the work described here we therefore used a different approach, suitable for a medium degree of oversizing. This is the "waveguide approach" in which the propagated wave is separated into the waveguide modes, between which conversions can take place.

$H$- and $E$-plane bends designed by the optical approach are shown in fig. 14. The incident wave, which may be represented by a "ray", falls upon a plane mirror at a certain angle and is reflected at the same angle towards the output port. However, some diffraction occurs, since in both waveguide arms a part of the wall has to be left out to permit the desired deflection of the ray. As the wall currents in the broad walls of a highly oversized waveguide are much higher than in the narrow walls, this diffraction is obviously more marked in the $E$-plane bend (on the left in fig. 14). In waveguide terms diffraction means excitation of higher-order modes; these belong to the $E_{m1}$ and $H_{m1}$ types for the $E$-plane bend and to the $H_{m0}$ types for the $H$-plane bend. Although the $E_{m1}$ and $H_{m1}$ modes can be absorbed by using the mode filters described above, the overall losses become so excessive (several decibels) that the $E$-plane bend designed by the optical method is not very attractive. On the other hand, the $H$-plane bend gives satisfactory results provided the waveguide is sufficiently oversized. An extensive theoretical analysis has been given by Dr. C. J. Bouwkamp of this laboratory \[18,\] For instance, the power loss due to mode conversion into the $H_{m0}$ modes has been shown to be less than 0.4 dB if the frequency exceeds eight times the cut-off frequency, in other words, if the waveguide is more than about five times oversized. However, it should be emphasized that the power loss in a complete system may be substantially higher because of trapped-mode resonances.

Since the optical approach has proved to be unsatisfactory in the case of an $E$-plane bend, a solution based on the waveguide principle has been sought. In the arrangement shown in fig. 15 the actual bend is made of flat waveguide, which is connected with the full-height oversized waveguides by means of two tapers. In the bend section there is some mode conversion into the $E_{m1}$ and $H_{m1}$ modes; but if the height of the flat waveguide does not exceed half the wavelength these modes are evanescent and do not appear at the ports. A bend of this type is usable from frequencies giving standard-size operation to those at which there is a high degree of overmoding, because the quantity determining the cut-off frequency of the desired mode, i.e. the width, does not change. The power loss, which is only of the order of a few tenths of a decibel, is mainly due to the wall losses in the flat guide.


Fig. 15. The 90° E-plane bend designed from a "waveguide approach" in which the wave is split up into waveguide modes. Conversion into $E_{x}$ and $H_{y}$ can occur, but these modes do not propagate in the flat guide provided $b_1 < \lambda/2$. In a design for use at $\lambda \geq 4$ mm the following dimensions were chosen: $a \times b = 22.86 \times 10.13$ mm, $r = 420$ mm; $b_1 = 2$ mm.

Fig. 17. Principle of a three-port directional coupler. At the left, transverse cross-section; on the right, longitudinal cross-section. Power from port 1 is divided in the ratio $(b - y)/y$ between the ports 2 and 3. Ports 2 and 3 are decoupled. 1 metallic sheet, 5 semi-transparent resistive sheet, 6 higher-order mode filter.

Fig. 16. Directional coupler designed from optical principles. $H$ is in the plane of drawing. At the dielectric sheet 5 the incident wave 1 is partly reflected (3), partly transmitted (4). There is some unwanted coupling 1 $\rightarrow$ 2 owing to diffraction, even without any sheet.

**Directional couplers**

A well-known optical design for a directional coupler is shown in fig. 16. It consists of an $H$-plane cross-junction of two waveguides, with a dielectric sheet placed in the diagonal plane of the cross. This sheet acts as a semi-transparent mirror and provides a directional coupling, as in optical interferometers. However, even without any sheet, a fundamental mode incident in

Fig. 18. Practical design of a 3-dB three-port directional coupler, from the principle of fig. 17.
one arm causes some radiation into the side arms owing to diffraction. The distribution of this deflected power as well as of the reflected and transmitted power among the different modes has been calculated by Katsenelenbaum\(^\text{[17]}\) and by Bouwkamp\(^\text{[18]}\). These diffraction phenomena give rise to certain errors in the performance of the directional coupler which decrease with increasing degree of overmoding. Special care has to be taken if the directional coupler is used close to the cut-off frequency of a higher-order \(H_{m0}\) mode which may give resonances in a system.

If there is no special reason for using a four-port directional coupler, and a three-port type of directional coupler with built-in load is sufficient, the following arrangement based on the waveguide principle is to be preferred\(^\text{[19]}\). A waveguide is divided into two parts by a septum parallel to the broad walls (fig. 17). This septum consists of two parts: a metal sheet and a resistive film which is semitransparent to microwaves. Power incident from the left (port 1) is divided between waveguides 2 and 3 in the ratio of their heights. This ratio determines the coupling factor of the directional coupler. Port 2 does not couple to port 3, because in two parallel guides separated by a homogeneous sheet, a wave propagating in one of the guides does not excite a wave propagating in the opposite direction in the other guide. The higher-order modes excited by the resistive sheet are attenuated by the mode filter at port 1. The resistive sheets of this filter and the resistive part of the septum are equivalent to the built-in load used in conventional standard-size directional couplers. A more practical form of oversized waveguide directional coupler is given in fig. 18. With this construction a directivity of about 50 dB has been obtained for a coupling of 3 dB (\(\lambda = 4\) mm, standard 3-cm waveguide). A great advantage of the present device is the frequency independence of the coupling, which is solely determined by the physical dimensions. Like the \(E\)-plane bend shown in fig. 15 the directional coupler is inherently broadband.

Some variable components

In principle the directional coupler described above

can also be used as a frequency-independent attenuator, if one arm is terminated in a matched load. In fig. 19 two such directional couplers are connected symmetrically in cascade, so that the input and output waveguides are full-height again. The device can easily be made into a variable attenuator by making the septum movable. As each of the directional couplers with one arm matched is a reciprocal two-port the total attenuation (in dB) is theoretically equal to twice the coupling (in dB) of each directional coupler; it is given by (cf. fig. 19):

\[
\text{attenuation} = 20 \log \left( \frac{b}{y} \right).
\]  

(11)

![Variable phase-shifter (line stretcher)](image)

Fig. 20. Variable phase shifter (line stretcher). The phase shift between the two ends of a waveguide is varied by varying the length of the line. By turning the knob the part 1 is made to slide in the transverse direction, thereby causing part 2 of the waveguide to slide in the longitudinal direction over the fixed part 3. The top and bottom of the waveguide are partly formed by 1. 4 is a thin metal sheet covering the hole of varying size between 2 and 3.

Measurements ($\lambda = 4$ mm, standard 3-cm waveguide) have indicated a minimum attenuation of 0.5 dB, and a maximum attenuation of 50 dB.

Fig. 20 shows a simple construction of a variable phase-shifter; some of the details are explained in the caption. This “line stretcher” can be used in all cases where the variation of its total length causes no difficulty.

A variable phase-shifter and a variable attenuator connected in cascade and terminated in a short-circuit constitute a variable impedance. Since in this case the output port need not be accessible, the arrangement can be simplified by using half of the attenuator shown in fig. 19. This gives the configuration of fig. 21 where the metallic sheet of the directional coupler has clearly become superfluous. The magnitude of the reflection coefficient can be varied by vertical displacement of the short-circuiting part together with the resistive sheet and the load: its value is simply given by $y/b$. The phase is varied by the line-stretcher described above, which is placed in front of the present device.

**Conclusions**

Now that we have described the particular features of several special components at some length, let us summarize some principal aspects of the design of single-mode oversized waveguide components. We have seen that the components designed from optical principles work satisfactorily only if the waveguide is sufficiently oversized. If, on the other hand, the waveguide is more than about ten times oversized, the tapers and mode filters — indispensable components for most applications — have to be inconveniently long. The quasi-optical components will therefore be usable only in a limited frequency interval, roughly characterized in practice by the limits “five times oversized” and “ten times oversized”, corresponding to an octave. On the other hand, some of the components described in this paper and designed from “waveguide” principles, work over a frequency decade, that is to say, between the frequencies corresponding to “not oversized” and “ten times oversized”.

Almost all components give rise to some mode conversion into unwanted modes. This phenomenon is annoying not only because of the loss in the power transmitted in the fundamental mode, but also because of possible interactions between the higher-order modes excited by different components and, in particular, the possible occurrence of trapped-mode resonances. On account of these effects mode filters have to be used which suppress or at least reduce these interactions and resonances. Since for the higher-order modes of class 1 \((E_{mn} \text{ and } H_{mn} \text{ modes with } n \neq 0)\) almost ideal filters can be constructed in the form of resistive sheets, the preferred components should be those which only give conversion into these modes. This is the case with the \(E\)-plane bend of fig. 15 and the directional coupler of fig. 18. These components have the common characteristic that all longitudinal cross-sections parallel to the narrow wall are identical. Furthermore, they are inherently broadband because the only limitation to the frequency of the dominant mode to be transmitted is a lower limit: the cut-off frequency determined by the width of the waveguide.

Tolerances in the form and dimensions of components have not been discussed in this article. Due attention has to be paid to precision manufacturing of the components and careful alignment in a system, as any deviation from ideal dimensions, any discontinuity at the flanges or inhomogeneity along the guide, can in principle give rise to higher-order modes.

Finally, let us enumerate some of the applications of oversized rectangular waveguide equipment:

1) Short-distance transmission, e.g. in radar aerial feeds.
2) Measurements of physical properties of materials such as dielectric constant, conductivity, surface impedance.
3) Simultaneous transmission of microwave signals of several frequencies covering a wide band (multiplex).
4) High-power transmission. This application, however, may be limited by breakdown or excessive rise of temperature in some of the components described in this paper.
5) It may also be possible to apply the oversized rectangular waveguide method to the submillimetre region, by scaling down the components.

In applications such as these the rectangular waveguide has proved to be superior to the circular waveguide operated in the \(H_{01}\) mode. A serious drawback of the \(H_{01}\) circular mode is that it is very difficult to design mode-launchers and filters, which means that its use is justified only if full advantage can be taken of the extremely low loss in this mode. Mode control in the circular guide is rather difficult, particularly since the \(H_{01}\) mode in this guide is degenerate with the \(E_{11}\) mode, so that mode conversion easily takes place.

Summary. The use of oversized waveguide for the transmission of millimetre waves has advantages over standard-size waveguide: the larger guide is easier to manufacture and to handle, the losses are smaller, and very broadband components can be made. For measuring purposes in particular, single-mode operation of the oversized waveguide is essential, as with multimode operation the field pattern is in practice unpredictable. If single-mode operation is to be achieved, mode control is necessary, i.e. measures have to be taken to prevent the conversion of the fundamental mode into higher-order modes, or to eliminate the effects of such conversion. An essential element for examining the mode conversion of components is the mode transducer, which converts a specific higher-order mode into the fundamental mode. Another indispensable element for mode control is the mode filter, which ideally absorbs all the higher modes without attenuating the fundamental mode. Several mode transducers and filters are described. Some of the oversized waveguide components discussed: tapers, \(E\)-plane bends, directional couplers, variable attenuators, phase shifters and impedances can be used over a 10 to 1 frequency band.