IDEAL W.G. TO COAX TRANSITIONS USING A F.B.M. MONOPOLE

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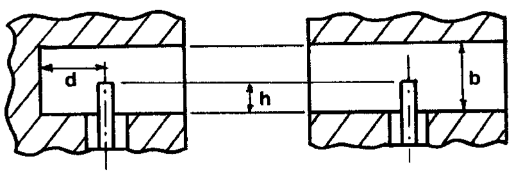
ABSTRACT

A w.g. - coax probe transition is in fact a monopole in a waveguide. We can match this, without tuning screws, over the full w.g. band for all types of w.g.: radial, circular TM_{01}, rectangular, single or double ridge and trough guide. Ideal rect. w.g. to coax transitions, right angle-, as well as end launch ones, have been realised: V.S.W.R. $\leq$1.02 for X-band with beadless coax connectors.

INTRODUCTION

Of the many configurations of w.g.-coax transitions, the probe type (Fig. 1a) is most popular due to its simple construction and rather good performance (V.S.W.R. $\leq$1.25) without special precautions. Attempts to realise ideal transitions (V.S.W.R. $\leq$1.02) failed, unless tuning screws were used. These perturb the reproducibility and even raise the costs. At any frequency in the band a perfect match can be obtained with the right combination of probe height (h) and - distance (d) from the plunger [1]. For which configuration can both be kept constant and still automatically full-band matched (f.b.m.)? For the symmetrical case (Fig. 1b), only h has to be kept constant; the straight rect. w.g. has been excited by a monopole. Having f.b.m. a monopole in free space first, we also learned how to match transitions from coax - via the monopole - to radial, to circular (TM_{01}) mode, to rectangular w.g., the trough guide and the single- and double-ridge w.g. They are all frequency dependent impedance steps which can now be f.b.m. [2].



a. asymm. w.g. b. symmetrical w.g.

Fig. 1 PROBE TYPE W.G.- COAX TRANSITIONS

I. A FULL-BAND MATCHED QUARTERWAVE MONOPOLE

A coaxial line protrudes with its inner conductor to a height h ($\cong\lambda/4$) above a metal ground plane, this way forming a monopole (Fig. 2a). Its equivalent circuit (Fig. 2b) is a series resonator with total impedance:

$$Z_M = R_M + j(\omega L_M - 1/\omega C_M)$$
$$= R_M(\omega) + jX_M(\omega) \qquad \text{(Fig. 2c)}$$

at resonance:

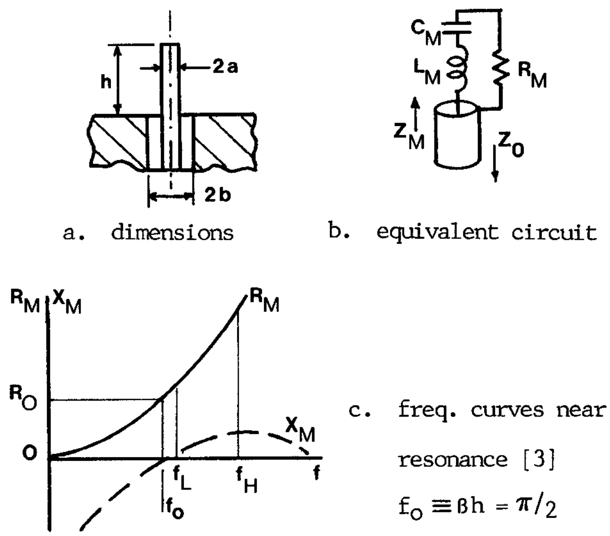
$$X_M = 0, \text{ so } Z_M = R_o \cong 36\,\Omega \text{ for } h \cong 0.23\,\lambda_o$$

From experiments [3], we deduced rather simple expressions:

$$R_M(\omega) = R_o \tan^2 \beta h/2$$

$$\text{and} \quad X_M(\omega) = X_o - X_{max} \sin 2\beta h$$

Being a rather frequency dependent impedance step with a small reactive part, towards the coax, it could be f.b.m. by a "reduced-quarterwave" transformer [2].



a. dimensions b. equivalent circuit

c. freq. curves near resonance [3] $f_o \cong \beta h = \pi/2$

Fig. 2 A QUARTERWAVE MONOPOLE

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The reduced-quarterwave transformer

It is well known that a transmission line of length l_1 and char. impedance Z_1 transforms an impedance Z_o into an impedance $Z_i = R_i + jX_i$ where (see Fig. 3a)

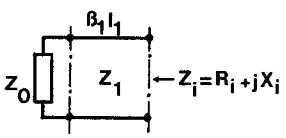
$$\frac{R_i}{Z_o} = \frac{1 + \tan^2 \beta_1 l_1}{1+(Z_o/Z_1)^2 \tan^2\beta_1 l_1} \quad \text{and}$$

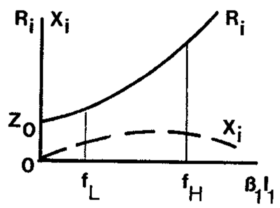
$$\frac{X_i}{Z_o} = \left(\frac{Z_1}{Z_o} - \frac{Z_o}{Z_1}\right) \frac{\tan \beta_1 l_1}{1+(Z_o/Z_1)^2 \tan^2 \beta_1 l_1}$$

For actual values $Z_o = 50\Omega$, $Z_1 = 72\Omega$ and $\beta_1 l_1 = \phi$ these expressions simplify to:

$$\frac{R_i}{Z_o} = 1 + \frac{2\tan^2 \phi/2}{1+\tan^4 \phi/2} \quad \text{and} \quad \frac{X_i}{Z_o} = 0.72 \frac{\sin 2\phi}{1+\cos^2\phi}$$

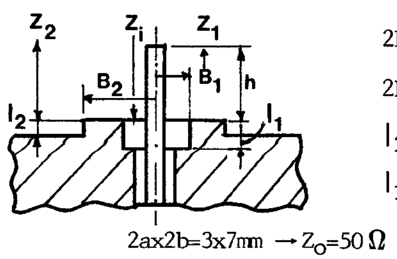
From Fig. 3b we see that their frequency curves are similar to those of the monopole (Fig. 2c), so a full-band match seems possible. If, finally, it proves difficult to obtain a perfect match over the whole frequency band, a radial "reduced-quarterwave" transformer might be added to the coaxial one (fig. 4). With these four independent parameters, an ideal match proved possible: the frequency curves to be matched are rather monotonous functions.





a. principle, $l_1 \leqslant \frac{\lambda_{min}}{8}$ b. freq. curves for $Z_1 = 71\Omega$, $l_1 = 3.5$mm

Fig. 3 THE REDUCED-QUARTERWAVE TRANSFORMER



$2B_2$ = 15 mm

$2B_1$ = 9.8 mm

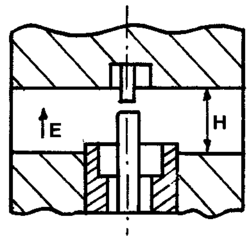
l_2 = 2 mm

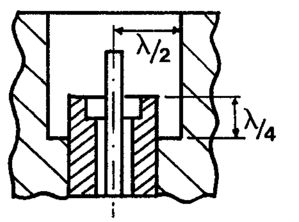
l_1 = 3.5 mm

Fig. 4 COAXIAL – & RADIAL RED. $\lambda/4$ TRANSFORMER (B_1, l_1 resp. l_2, B_2) IN F.B.M. MONOPOLE

II THE F.B.M. MONOPOLE IN DIFFERENT SYMMETRICAL WAVEGUIDES

Once the monopole is matched we can apply it in many w.g. types. Looking into each waveguide, the coaxial line sees a different frequency dependent impedance. Its real part is mainly determined by the transmitted power – its far field – so by h and its coaxial transformer; its imaginary part is representative for the higher-order modes – its near field – and can best be matched locally, so with the radial transformer or other local matching elements. We will start with the simplest ones: the rotation symmetrical waveguides.





a. radial guide b. TM_{01} circular w.g.

Fig. 5 A F.B.M. MONOPOLE IN ROTATION SYMM. W.G.

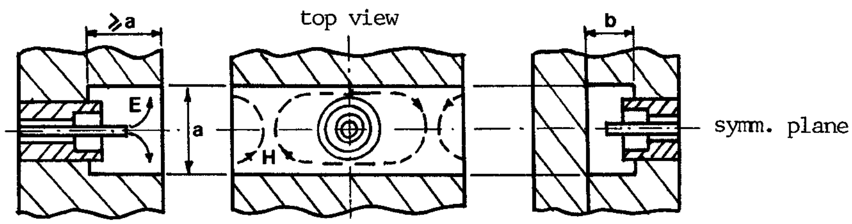
The radial waveguide.

Screening-off the monopole above a parallel metal plane at height ($H \leq \lambda/2$), a radial w.g. has been obtained (Fig. 5a). The load for the monopole has changed, but matching can be achieved again by changing h, l_1 and Z_1. If the antenna top comes too close to the top plate, a blind hole must be applied, eventually to be used for matching as well. After optimising l_1, Z_1 and l_2, Z_2 an ideal w.g. transition (V.S.W.R. $\leq$1.02 for X-Band) could be realised; very useful as a precision coax mm load and for power combining [4].

The TM_{01} circular waveguide.

The simplest way to excite a pure TM_{01} mode in circular w.g. is by a coaxial line (Fig. 5b). In order to prevent high reflection from the w.g. walls, located about $\lambda/2$ from the monopole, a high series impedance has been inserted: the monopole was placed a quarterwave away from the shorted w.g. end. Only minor corrections were needed for h and the "reduced" transformers for full-band matching (V.S.W.R. $\leq$ 1.05).

We continue with the straight symm. waveguides:



c. trough guide

d. rectangular w.g.

Fig. 5 A F.B.M. MONOPOLE IN STRAIGHT SYMM. W.G.

The trough guide.

This guide can be obtained by symmetrically providing a monopole with metal sidewalls (Fig. 5c). For a distance "a" apart, the magnetic field lines are squeezed as in a rectangular w.g. Matching can be achieved with the coaxial - and radial transformer.

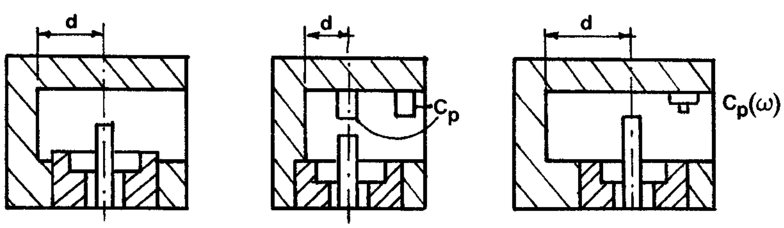
The rectangular waveguide

Closing the trough with a top plate at a height b, we obtain a coax-to-symmetrical rectangular w.g. (cross-section a x b) transition (Fig. 5d). In a similar way as above the coaxial- and radial-transformer provide the matching, correcting its dimensions for the change in configuration.

III IDEAL RECT. W.G.-COAX TRANSITIONS

The right angle type

Usual w.g. to coax transitions are asymmetrical (Fig. 6), one w.g. arm is short-circuited by a plunger, a distance d away from the monopole. Above we have seen how to obtain h constant. Now we must try to achieve also d constant for full-band matching. First h, l_1 and Z_1 must be such that matching at any frequency in the band can be obtained by varying only d. Now for f.b.m., d must be kept constant, which can be done for d = $\lambda g/4$ for the lowest frequency (f_L) in the band, the medium one (f_M) or the highest one (f_H). The simplest and least critical one is at f_M. The short-circuited w.g. of length d represents a parallel resonator in the plane of the antenna, being resonant for f_M. Then reflections at f_L and f_H are already rather low, but with the right dimensions of h, l_1, Z_1, l_2 and Z_2 (fig. 4) a perfect f.b.m. (V.S.W.R. $\leq$ 1.02) could be obtained (Fig. 6a). The shortest, but most critical choice for d is $\lambda g/4$ at f_H, so a shunt inductance to the monopole over the band. Its reflection can be compensated by a shunt capacitance with the same frequency curve, so a double cap. stub C_p (Fig. 6b). As one stub is close to the antenna top, h has to be corrected, which is not attractive. A more independent method is by chosing d = $\lambda g/4$ at f_L, so a shunt capacitance to the monopole over the band, which reflection can be compensated by a special capacitive stub a quarterwave in front of it (Fig. 6c).



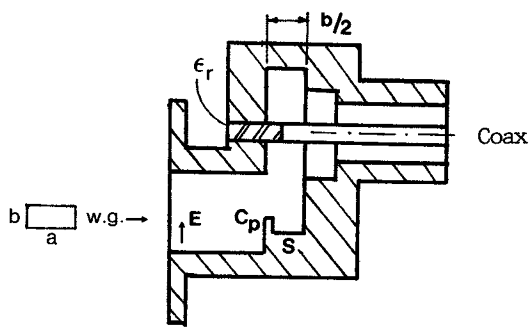
d = $\lambda g/4$ for : a. f_M($\simeq$10GHz) b. f_H($\simeq$12.4GHz) c. f_L($\simeq$8.2GHz)

Fig. 6 F.B.M. COAX-RECT. W.G. TRANSITIONS WITH DIFFERENT d's

The end-launch type.

A f.b.m. monopole is used here as well, so a 90° corner is needed, either in coax, in waveguide or in the antenna. A waveguide corner was preferred, not only in view of mm waves, but at the same time for easy matching of the transition (fig. 7).

The "reduced-quarterwave" transformer has been realised by the lower waveguide (b/2). The reflection of the corner and the impedance step could be so reduced - by way of a small step S plus a shunt capacitance C_p in the form of a metal strip - that it forms the right matching element for the transition. This way an extremely short, ideal (V.S.W.R. ≤ 1.02) transition has been realised.



M.E.: S = step; C_p = parallel cap.

ϵ_r = dielectric support for inner conductor of coax

Fig. 7 END-LAUNCH COAX-RECT. W.G. TRANSITION

CONCLUSION

In a systematic way ideal transitions from coax to different types of waveguides can now be realised without the need for tuning screws. They are all based on the full-band matching of a monopole in the particular w.g. Consequently, identical pairs can easily be realised, ideal for bridge measurements.

ACKNOWLEDGEMENT

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REFERENCES

1. R.E. Collin , "Field theory of guided waves", McGraw Hill, N.Y. 1960, Chapter 7.
2. F.C. De Ronde, "An octave-wide matched impedance step and - quarterwave transformer", Int. Microwave Symp. MTT - Baltimore 1986, Digest pp 151-154.
3. R.W.P. King, "The theory of linear antennas", II.38, Harvard Univ. Press, Cambridge, Mass. 1956.
4. R.C. Allison, R.L. Eisenhart, P.T. Greiling, "A matched coaxial-radial transmission line junction", Int. Microwave Symp. MTT 1978, Digest pp 44-46.