# Broad-Band Coaxial Choked Coupling Design*

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*Summary*—**Equations and curves are presented to predict the frequency bandwidth of coaxial choke couplings in terms of the choke parameters. Choke couplings discussed are those applicable to rotary joints and dc isolation units.**

## Introduction

COAXIAL choke-type rotary joint designs have been discussed by Ragan,[1] and many joints have been built following his presentation. Recently, Muehe[2] discussed a method to widen the bandwidth of coaxial choke-type rotary joints by reducing the characteristic admittance of the transmission line for a quarter wavelength on each side of the chokes. Muehe's discussion was based on the analogous case of broadbanding short-circuited quarter wavelength stubs in parallel with the transmission line, by changing the characteristic impedance of the line on each side of the stub for a distance of one-quarter wavelength.

The broadbanding of coaxial choke couplings under the present discussion is not based on a change in transmission line impedance, but is based on an extension of the conventional methods. As outlined by Ragan, broadbanding of choke couplings may be accomplished by displacing the outer and inner conductor chokes along the transmission line by one-quarter wavelength. The purpose of this paper is to present general equations and curves relating to the VSWR, to the characteristic impedance of the choke sections, and to the spacing of the two chokes. From these curves, one can predict the bandwidth of a rotary joint design.

## Equations for Choke Coupling

A conventional coaxial rotary joint is shown in Fig. 1. To prevent radiation losses due to the outer conductor choke and to provide a means of placing a bearing at a low current point, the external choke section of characteristic impedance $Z_{03}$, is added. In most practical cases, the characteristic impedance $Z_{03}$ is made as high as possible and is usually much greater than the characteristic impedance $Z_{02}$ of the outer conductor choke.

For simplification in this analysis, $Z_{03}$ is assumed to be infinite. Also, both the choke sections are $\lambda/4$ long at the center frequency, and their characteristic impedances $Z_{01}$ and $Z_{02}$ are assumed to be equal; thus, the choke input impedances are equal, or $Z_1 = Z_2$. The characteristic impedance, $Z_0$, of the transmission line is normalized to 1.

The *ABCD* matrix of the two chokes displaced by the length $l$ along a lossless transmission line is

$$\begin{bmatrix} 1 & -jZ_{01}\cot\beta l_1 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} \cot\beta l & j\sin\beta l \\ j\sin\beta l & \cos\beta l \end{bmatrix} \cdot \begin{bmatrix} 1 & -jZ_{01}\cot\beta l_1 \\ 0 & 1 \end{bmatrix}, \quad (1)$$

where $\beta l_1$ is the electrical length of the choke sections and $\beta l$ is the electrical spacing between the two chokes. The final matrix when multiplied through is

$$\begin{bmatrix} \cos\beta l + Z_{01}\cot\beta l_1 \sin\beta l & -2jZ_{01}\cot\beta l_1\cos\beta l - jZ_{01}^2\cot^2\beta l_1\sin\beta l + j\sin\beta l \\ j\sin\beta l & \cos\beta l + Z_{01}\cot\beta l_1\sin\beta l \end{bmatrix}. \quad (2)$$

In addition to facilitating the design of rotary joints, the information presented here can be utilized to build wideband dc isolation units. Wideband dc isolation units have been developed[3] using the design described; dc isolation units are required whenever blocking of dc on both the inner and outer conductors is desired.

The insertion loss is given by[4]

$$\begin{aligned} L &= 10\log_{10}\{1 + 1/4[(A-D)^2 - (B-C)^2]\}, \\ &= 10\log_{10}\{1 - 1/4[-j(2Z_{01}\cot\beta l_1\cos\beta l \\ &\qquad + Z_{01}^2\cot^2\beta l_1\sin\beta l)]^2\}, \\ &= 10\log_{10}\left\{1 + \frac{|K|^2}{4}\right\}. \end{aligned} \quad (3)$$

Next, let

$$\beta l_1 = \frac{\pi}{2} + \phi \quad \text{and} \quad \beta l = n\beta l_1 = n\left(\frac{\pi}{2} + \phi\right).$$

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[1] G. L. Ragan, "Microwave Transmission Circuits," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 9, p. 407; 1948.

[2] C. E. Muehe, "Quarter-wave compensation of resonant discontinuities," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 296–297; April, 1959.

[3] By Ramo-Wooldridge, a division of Thompson Ramo Wooldridge Inc., Los Angeles, Calif.

[4] R. M. Fano and A. W. Lawson, "Microwave Transmission Circuits," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 9, p. 551; 1948.

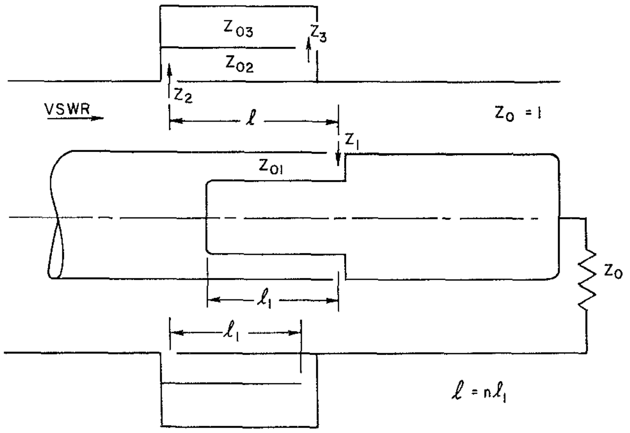


Fig. 1—Coaxial transmission line with choke couplings used as a rotary joint.

Then $|K|$ is written as

$$|K| = 2Z_{01} \tan \phi \cos n\left(\frac{\pi}{2} + \phi\right) - Z_{01}^2 \tan^2 \phi \sin n\left(\frac{\pi}{2} + \phi\right), \quad (4)$$

where $\phi$ = the difference in electrical length from the center frequency choke length of $\pi/2$. The relation of $|K|$ to the coaxial transmission line voltage-standing-wave ratio, VSWR, is

$$|K| = \frac{VSWR - 1}{\sqrt{VSWR}}. \quad (5)$$

Bandwidth is arbitrarily defined as the ratio of the two frequencies for which the transmission line VSWR is 1.1:1 (insertion loss less than 0.01 db) or in symbolic form, the bandwidth is

$$\frac{f_2}{f_1} = \frac{90 + |\phi_2|}{90 - |\phi_1|}. \quad (6)$$

## Discussion of Curves

A graph of the input VSWR to the transmission line is shown in Fig. 2 for the case when $Z_{01} = 0.0324$ for various conditions of $n$. When $n = 0$, the input terminals of the chokes are located on the same transverse plane. The VSWR is calculated from

$$|K|_{n=0} = 2Z_{01} \tan \phi. \quad (7)$$

When the chokes are displaced by $\lambda/4$ at the center frequency, then $n = 1$, or (4) is reduced to

$$|K|_{n=1} = (2Z_{01} + Z_{01}^2) \tan \phi \sin \phi. \quad (8)$$

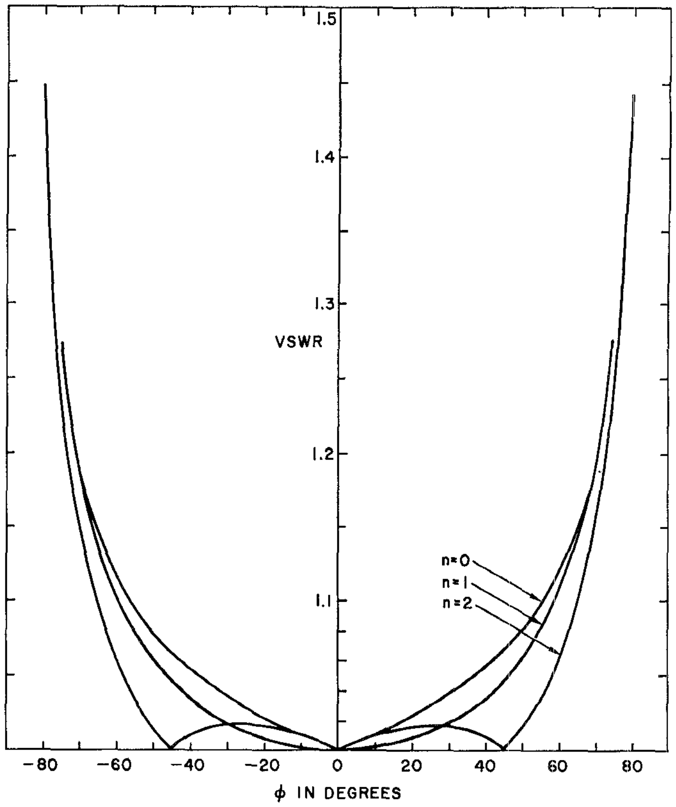


Fig. 2—Input VSWR to coaxial transmission line with choke couplings for various values of $n$. Choke impedance, $Z_{01} = 0.0324$ (normalized to unity). Characteristic impedance $Z_{03}$ is assumed to be infinite.

When $n = 1$, a zero derivative exists at the origin, a condition considered to be the maximally flat case.

When the input choke terminals are separated by $\lambda/2$ at the center frequency ($n = 2$), the $n = 2$ curve shows a wider bandwidth compared to the $n = 1$ condition. For the $n = 2$ condition, $|K|$ is written as

$$|K|_{n=2} = 2Z_{01} \tan \phi \cos^2 \phi - 2(Z_{01} + Z_{01}^2) \tan \phi \sin^2 \phi. \quad (9)$$

When the chokes are separated by $\lambda/4$ at the center frequency, or $n = 1$, the frequency bandwidth ratio vs $Z_{01}$ is plotted in Fig. 3, where the band edge limits were determined by a voltage-standing-wave ratio of 1.1:1. Fig. 4(a) is a curve of frequency bandwidth ratio vs $Z_{01}$ for $n = 2$. The peak voltage-standing-wave-ratio within the band limits is plotted in Fig. 4(b).

Where $Z_{01} = 0.0324$ and $n = 2$, the bandwidth for a VSWR of less than 1.1:1 is 6.2:1. With the same choke impedance except that $n = 2.67$ ($\beta l = 240°$ at the center frequency), a still wider frequency bandwidth of 8.15:1 is theoretically feasible as illustrated by the solid line curve of Fig. 5. Note that the curve is unsymmetrical and the peak within the band is slightly higher than for the case when $n = 2$. In actual experience, the slightly higher peak should not produce any detrimental ef-

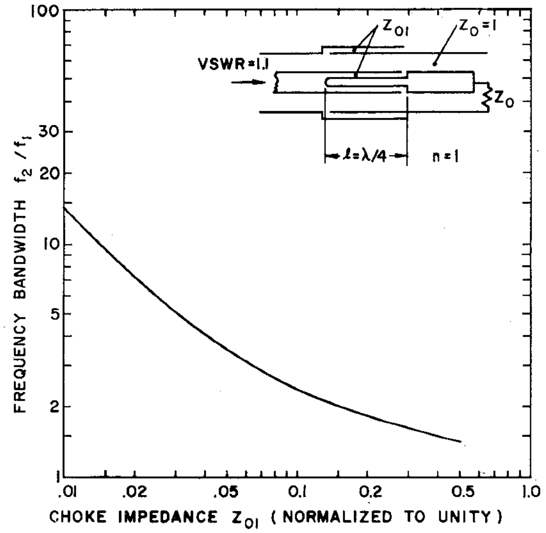


Fig. 3—Frequency bandwidth curve of coaxial transmission line with choke couplings when chokes are spaced $\lambda/4$ at the center frequency. Bandwidth determined by the transmission line VSWR of 1.1:1. Characteristic impedance $Z_{03}$ is assumed to be infinite.

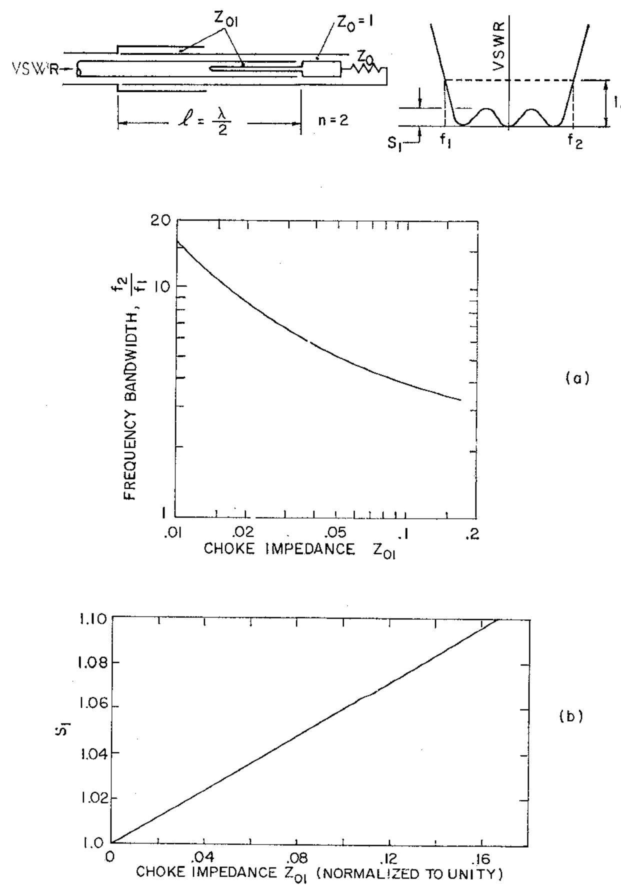


Fig. 4—Response curves for coaxial transmission line with choke couplings when chokes are spaced $\lambda/2$ apart at the center frequency. (a) Frequency bandwidth determined by the transmission line VSWR of 1.1:1. (b) The value of the peak VSWR within the band limits. Characteristic impedance $Z_{03}$ is assumed to be infinite.

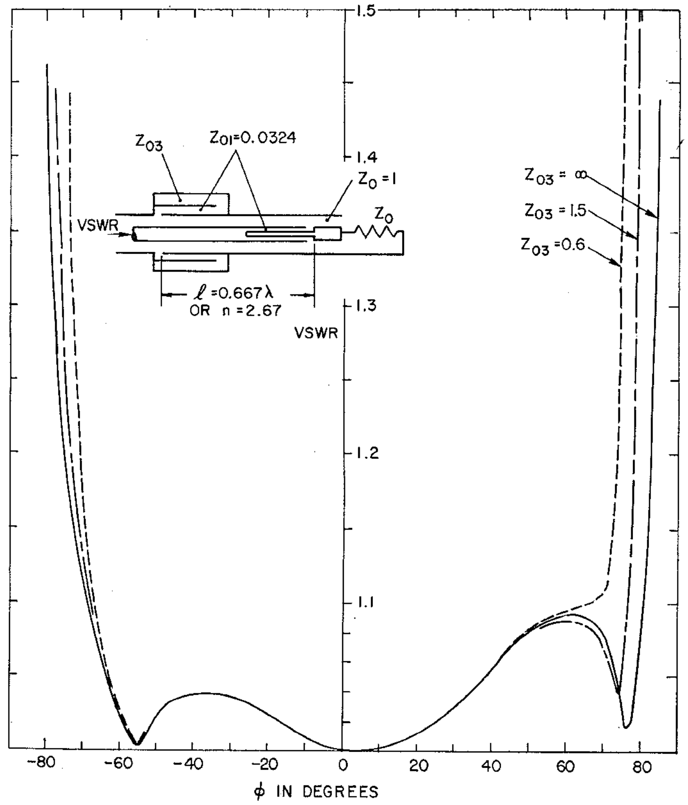


Fig. 5—Input VSWR curve of coaxial transmission line with choke couplings when chokes are spaced $0.667\lambda$ apart for $Z_{01} = 0.0324$ (normalized to unity).

fects. With any given value of $Z_{01}$ an optimum spacing between chokes can be determined to give the widest frequency bandwidth.

## Other Considerations

For extremely wide bandwidth choke couplings, the finite value of $Z_{03}$ should be considered. The solid line curve of Fig. 5 represents the condition when the external choke impedance $Z_{03}$ is infinite while the broken line curves represent two finite values of $Z_{03}$.

The dashed line curve of Fig. 5 shows the reduction in bandwidth if the normalized characteristic impedance of $Z_{03}$ is 0.6. For a normalized characteristic impedance $Z_{03}$ of 1.5, the dash-dot curve indicates an improvement in the VSWR response. Note that there is no major increase in VSWR due to the finite value of $Z_{03}$ provided $|\phi| < 70°$. Usually only for extremely wide bandwidth choke coupling designs (when $|\phi| > 70°$) is it necessary to calculate the effects of $Z_{03}$ on the VSWR response.

If the external diameter is a limiting factor that prevents a large value of $Z_{03}$ from being selected, an effective increase in the external choke impedance can be obtained by making it $\lambda/2$ long at the midfrequency as illustrated in Fig. 6. Now the impedance $Z_3 + Z_4$ terminates the outer conductor choke $Z_{02}$, instead of only $Z_3$.

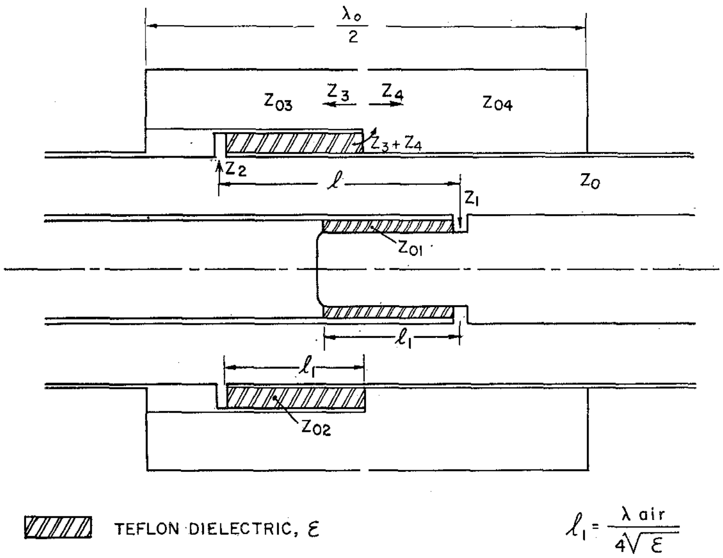


Fig. 6—Broadband coaxial rotary joint of dc isolation unit design using choke couplings.

In a practical design, the slight reduction in bandwidth due to the finite impedance of the external choke can be compensated by lowering the characteristic impedance $Z_{02}$. In a practical case, $Z_{02}$ can be made smaller than $Z_{01}$.

Basically, the wide bandwidth of choke couplings is obtained by the use of low values of the choke's characteristic impedances, $Z_{01}$ and $Z_{02}$. By the use of teflon insulation, choke characteristic impedance of less than 1.5 ohm is readily obtained. The increase in dissipation losses because of the use of teflon dielectric is compensated for by the shorter physical length[5] of the choke that is required.

Fig. 7 is a photograph of a dc isolation unit built[3] for a frequency range of 70 mc to 850 mc (VSWR less than 1.5) using the design described herewith. Performance data of the dc isolation unit is illustrated in Fig. 8, showing both the calculated and measured VSWR response. The measured insertion loss was found equal to the mismatch loss.

## Conclusion

The curves (Figs. 3 and 4) and the analysis should be helpful for the design of wide-band coaxial rotary joints or dc isolation units. Prediction of the final performance for various choices of parameters is possible.

## Appendix

### Choke Losses Due to Teflon Dielectric

To demonstrate the effects of the choke losses due to the use of teflon dielectric, the following relations are presented. The input resistance to a resonant transmission line with a relative dielectric constant of unity, is

[5] See Appendix.

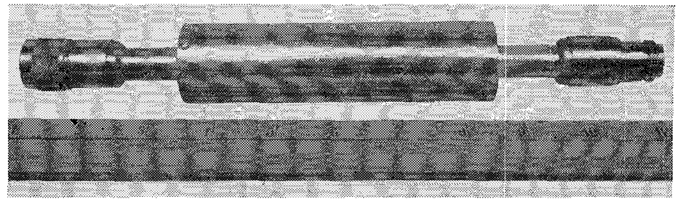

Fig. 7—DC isolation unit built for 70 mc to 850 mc operation (VSWR less than 1.5) using the choke coupling design described.

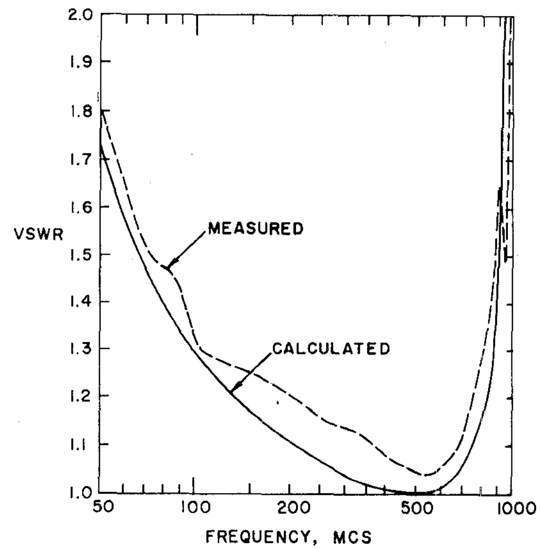


Fig. 8—Measured and calculated VSWR curves of the dc isolation unit of Fig. 7.

expressed as

$$R_{\text{in}_{\epsilon=1}} = Z_0 \alpha l = \frac{Rl}{2} + \frac{GZ_0{}^2 l}{2}, \tag{10}$$

where

$$\alpha = \frac{R}{2Z_0} + \frac{GZ_0}{2}.$$

Assuming that the dielectric constant is increased to $\epsilon_r$, the characteristic impedance $Z_0$ and length $l$, are both reduced by $1/\sqrt{\epsilon_r}$. A dielectric-filled resonant line, assuming the same dielectric dissipation factor of equation (10) will have the input resistance

$$R_{\text{in}_{\epsilon=\epsilon_r}} = \frac{Rl}{2\sqrt{\epsilon_r}} + \frac{GZ_0{}^2 l}{2\epsilon_r{}^{3/2}}. \tag{11}$$

Note that the copper loss term is reduced by $1/\sqrt{\epsilon_r}$. Furthermore, the dielectric loss term is reduced by a factor of $1/\epsilon_r{}^{3/2}$.

In actuality, the dielectric loss is increased from zero to a finite value when converting an air dielectric filled choke to a teflon dielectric filled choke. However, the loss tangent of teflon is very small, $Z_0$ is small, and the dielectric loss term is also reduced by $1/\epsilon_r{}^{3/2}$. Thus, in a practical case, instead of increased dissipation losses due to the use of teflon dielectric, the losses are reduced.