Octave Bandwidth Tunnel-Diode Amplifier*

J. H. LEPOFF†, SENIOR MEMBER, IEEE, AND G. J. WHEELER†, SENIOR MEMBER, IEEE

Summary—**Two tunnel-diode amplifiers for the frequency ranges 2 to 3 Gc and 3 to 4 Gc are combined with hybrids and low-pass filters to provide an octave bandwidth amplifier. The amplifier design is based on band-pass filter techniques. The gain exceeds 10 db over this band with noise figure better than 4 db.**

Introduction

Tunnel-diode amplifiers are limited in bandwidth by stability problems and by the bandwidth of available circulators. The stability problem is solved by using an amplifier circuit with steep gain skirts. Then, as the circulator reflections rise outside the operating frequency band, the gain drops below the level needed to sustain oscillations. The bandwidth limitation of available circulators is overcome by combining two half-octave amplifiers in a hybrid-filter circuit to get full-octave amplification.

Filter-Type Amplifier Design

The use of prototype low-pass filters in the design of broad-band parametric amplifiers has been discussed by Matthaei[1] and by Mayer.[2] The technique is also suitable for the design of tunnel-diode amplifiers. A Chebyshev band-pass filter is used to determine the parameters of the amplifier. This filter is based on a low-pass prototype whose element values are available in tables for various values of ripple amplitude of the pass-band insertion loss.

Fig. 1 is an equivalent circuit of the amplifier. The load is $-R_D$, the diode negative resistance. The generator is R_g, the circulator impedance. In the equivalent circuit, the diode capacitance C_D has been included in B_1. Since distributed elements are used, the shunt element B_1 consists of C_D resonated by a length of shorted transmission line in shunt, while the series element X_2 is a half wavelength of shorted transmission line. The low-pass prototype shown in Fig. 2 has element values corresponding to 0.01-db ripple and minimum reflection loss.[3]

To obtain the parameters for Fig. 1, one must first evaluate the shunt capacitance C and series inductance

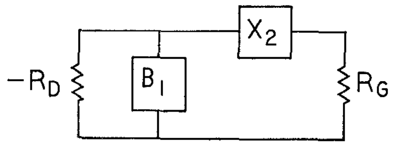

Fig. 1—Amplifier equivalent circuit.

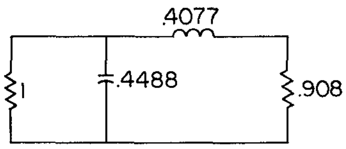

Fig. 2—Low-pass prototype filter.

L of the lumped element band-pass filter based on the low-pass prototype shown in Fig. 2.[4] The capacitance is divided by the bandwidth W and by the ratio of generator resistance to prototype load resistance 50/0.908. The inductance is divided by bandwidth and multiplied by the resistance ratio. For 1-Gc bandwidth, the capacitance is thus $8.15/2\pi$ pf and the inductance is $22.44/2\pi$ nh.

The amplifier circuit may now be evaluated by setting the slope parameters of its two elements equal to the slope parameters of the lumped element band-pass filter. The slope parameter of the shunt element is defined by $b_1 = (\omega_0/2)(dB_1/d\omega)$ with the derivative evaluated at ω_0, the geometric mean frequency. For the lumped element filter, $b_1 = \omega_0 C = 8.15 f_0 \times 10^{-3}$. For the lower band, $f_0 = 2.45$ Gc and $b_1 = 0.02$. The shunt element of the amplifier involves two unknowns, the characteristic impedance Z_{01} or admittance Y_{01} of the shunt stub and its length. The two necessary equations are the resonance equation and the slope parameter equation.

$$B_1 = 0, \tag{1}$$

$$\frac{\omega_0}{2} \frac{dB}{d\omega}\bigg|_{\omega=\omega_0} = 0.02, \tag{2}$$

or, using θ_{10} for electrical stub length at center frequency,

$$\omega_0 C_D - Y_{01} \cot \theta_{10} = 0 \tag{3}$$

and

$$\frac{\omega_0 C_D}{2} + \frac{Y_{01}\theta_{10} \csc^2 \theta_{10}}{2} = 0.02. \tag{4}$$

* Received May 20, 1963; revised manuscript received August 19, 1963. The work reported here was performed for the U. S. Army Electronics Research and Development Laboratory, Contract DA 36-039 AMC-0008(E), Electronic Defense Laboratories, Sylvania Electronic Systems—West, Mountain View, Calif.

† Electronic Defense Laboratories, Sylvania Electronic Systems—West, Mountain View, Calif.

[1] G. L. Matthaei, "A study of the optimum design of wide-band parametric amplifiers and up-converters," IRE Trans. on Microwave Theory and Techniques, vol. 9, pp. 23–27; January, 1961.

[2] A. Mayer, "Wideband single-diode parametric amplifiers," *Microwave J.*, vol. 5, pp. 99–106; February, 1962.

[3] G. L. Matthaei, *et al.*, "Design Criteria for Microwave Filters and Coupling Structures," Stanford Research Institute, Menlo Park, Calif., Project 2326 Final Rept., p. 202; January, 1961.

[4] L. Weinberg, "Network Design by Use of Modern Synthesis Techniques and Tables," Hughes Aircraft Co., Culver City, Calif., Tech. Memorandum 427, p. 18; April, 1956.

Eliminating Y_{01} gives

$$\frac{\sin 2\theta_{10}}{2\theta_{10}} = \frac{\omega_0 C_D}{0.04 - \omega_0 C_D} \cdot \quad (5)$$

Before solving, it is interesting to note that this equation defines a maximum value of C_D. The function $\sin X/X$ has a maximum value of unity, so

$$\omega_0 C_D = 0.04 - \omega_0 C_D \quad (6)$$

and

$$C_D < \frac{0.02}{\omega_0}$$

or, in general,

$$C_D < \frac{b_1}{\omega_0} \cdot \quad (7)$$

This may be considered a constraint on the bandwidth imposed by the diode, because b_1 is proportional to C which is inversely proportional to bandwidth. For this amplifier, the maximum C_D is 1.3 pf.

The design is less critical if C_D is not close to this maximum value. However, a 1.2-pf diode was available so the circuit was designed for it. The solution is $\theta_{10} = 26.8°$. Substituting in one of the equations gives a characteristic impedance Z_{01} of 107 Ω.

The design of the series resonator is straightforward. The resonance condition establishes the length as a half wavelength at center frequency or $\theta_{20} = \pi$. The series reactance is $Z_{02} \tan \theta_2$. The slope parameter, defined as $x_2 = (\omega_0/2)(dX_2/d\omega)$ evaluated at center frequency, is $\pi Z_{02}/2$. For the lumped element band-pass filter, $x_2 = \omega_0 L_2 = 22.44 f_0 = 55$ for the 2- to 3-Gc amplifier. Equating the two slope parameters shows that $Z_{02} = 2x_2/\pi = 35$ Ω.

In a similar manner, an amplifier for the 3- to 4-Gc frequency range was designed. The computed gain curves for these amplifiers are shown in Fig. 3. The amplifier bandwidth is considerably wider than the filter bandwidth. The power gain ratio equals

$$\frac{(R_G - R)^2 + X^2}{(R_G + R)^2 + X^2}$$

where R and X are the real and imaginary parts of the input impedance seen by the circulator. The high gain peaks occur when $R_G + R = 0$ and $X = 0$ at nearly the same frequencies. This is shown in Fig. 4 where R and X for the 2- to 3-Gc amplifier are plotted as functions of frequency. The gain peaks may be reduced by modifying the circuit parameters to separate the real and imaginary zeros. Reducing Z_{02}, for example, separates the imaginary zero frequencies with no effect on the real part of the input impedance. In this particular amplifier, the gain peaks could be reduced by using a smaller capacitance diode as shown in Fig. 3.

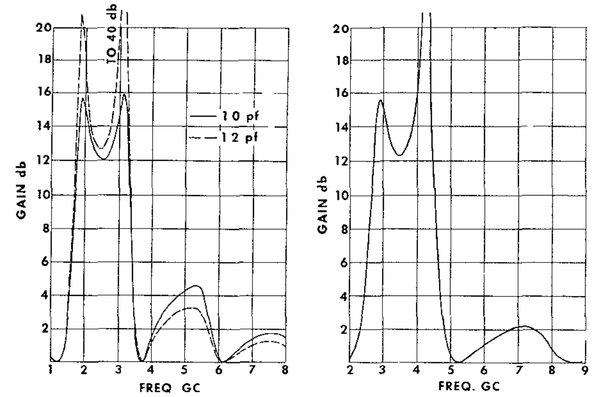

Fig. 3—Computed gain of Chebyshev amplifiers.

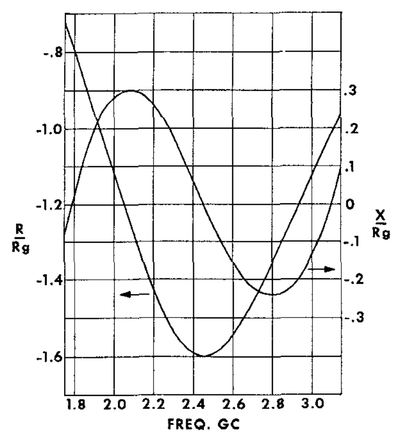

Fig. 4—Impedance of 2- to 3-Gc Chebyshev amplifier.

A more straightforward method of designing filter-type amplifiers uses the same circuit without a low-pass prototype to determine the element values. Instead, the imaginary part of the input impedance is made equal to zero at the band edges and the real part is made equal to the appropriate value to give the desired gain. For example, the shunt resonator for an amplifier with 11-db gain at center frequency and at band edges is designed by the equations

$$\frac{\tan \frac{f_1}{f_2}\theta_{12}}{\tan \theta_{12}} = \frac{f_2 C - 2.64}{f_1 C + 2.64}$$

and $Y_{01} = (0.00628C - 0.01656) \tan \theta_{12}$ with f_1 and f_2 the lower and upper band edge frequencies in gigacycles and C the diode capacitance in picofarads. The angle θ_{12} is the electrical length of the shunt stub at f_1.

The details are given in the Appendix. The first equation limits the diode capacitance between a maximum of 2.64 pf and a minimum of $2.64/f_2$ pf. Fig. 5 shows the variation of shunt resonator length and characteristic

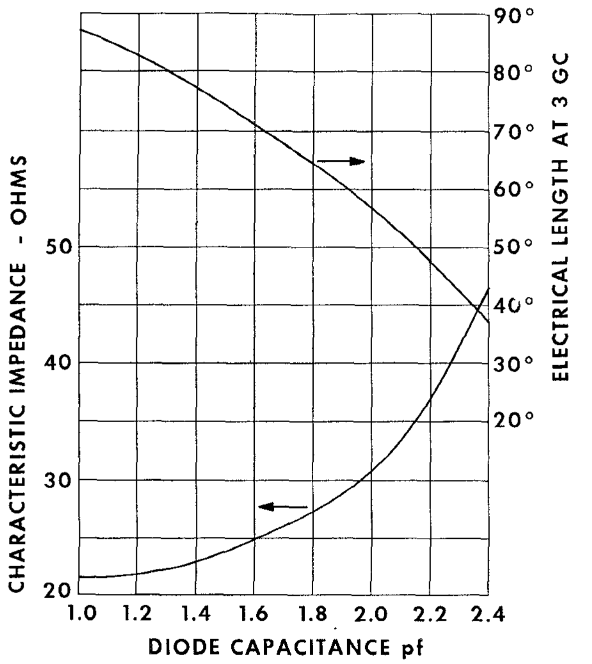

Fig. 5—Shunt resonator design for 2- to 3-Gc bandwidth.

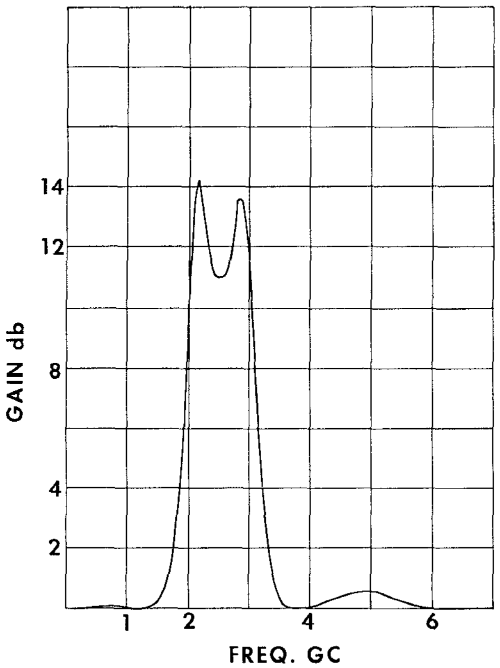

Fig. 6—Calculated gain response of tunnel-diode amplifier.

impedance with diode capacitance for the 2- to 3-Gc band.

The equations for the series resonator are

$$\theta = \frac{f_2 - f_1}{f_2 + f_1} 180°$$

and

$$Z_{02} = \frac{41.4}{\tan \theta} \cdot$$

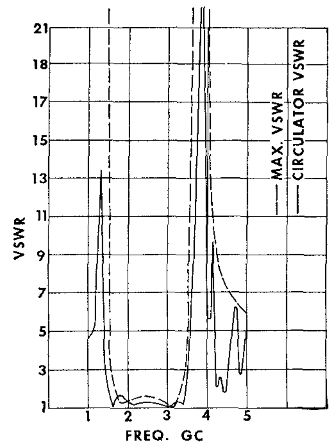

Fig. 7—A stability criterion for a tunnel-diode amplifier.

where Z_{02} is the characteristic impedance of the series resonator and θ is the difference between 180° and its electrical length at the band edges. For the 2- to 3-Gc band, $\theta = 36°$ and $Z_{02} = 57.1\ \Omega$. Fig. 6 shows the calculated gain response of the amplifier described above with a 50-Ω circulator and a 1.3-pf diode.

Stability

A condition for stability is that the loop impedance should not vanish in the right half of the complex frequency plane. It has been shown[5] that this leads to a sufficient condition for stability,

$$|G|\,|\rho| < 1,$$

where G is the voltage gain of a stable amplifier with an ideal resistive load and ρ is the reflection coefficient of the actual load such as a circulator. The calculated gain of an amplifier may be used with this condition to establish a maximum VSWR for the circulator. If the circulator VSWR is less than this maximum, the circuit is stable. If the circulator VSWR is greater, the phase of the reflection determines whether or not the circuit is stable. Fig. 7 shows the computed VSWR limit for the 2- to 3-Gc tunnel diode amplifier described earlier. The lower curve is the measured VSWR of the circulator used in this band. The phase of the reflection was adjusted experimentally to make the amplifier stable despite the violation of the stability criterion in a narrow frequency range.

This stability criterion is actually not applicable at the frequencies where the maximum VSWR reaches a peak. At these frequencies, the series resonator is an open circuit. The circulator is disconnected from the diode so its impedance has no effect on stability. If the shunt element were resonant when the series element is open, the circuit would be unstable. In practical circuits this is not the case.

[5] B. Hennock and Y. Kvaerna, "Stability criteria for tunnel-diode amplifiers," IRE Trans. on Microwave Theory and Techniques, vol. 10, pp. 397–398; September, 1962.

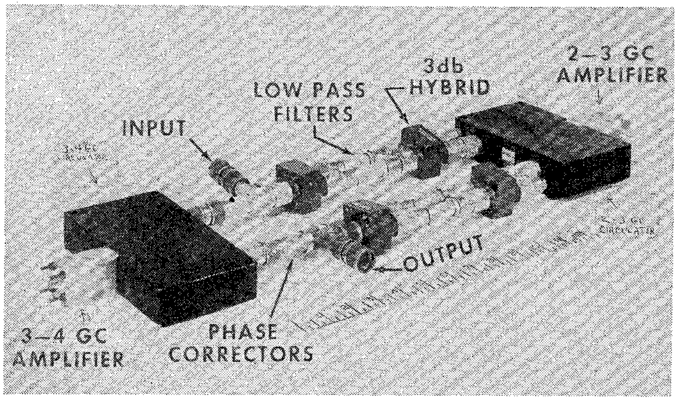

Fig. 8—Octave amplifier.

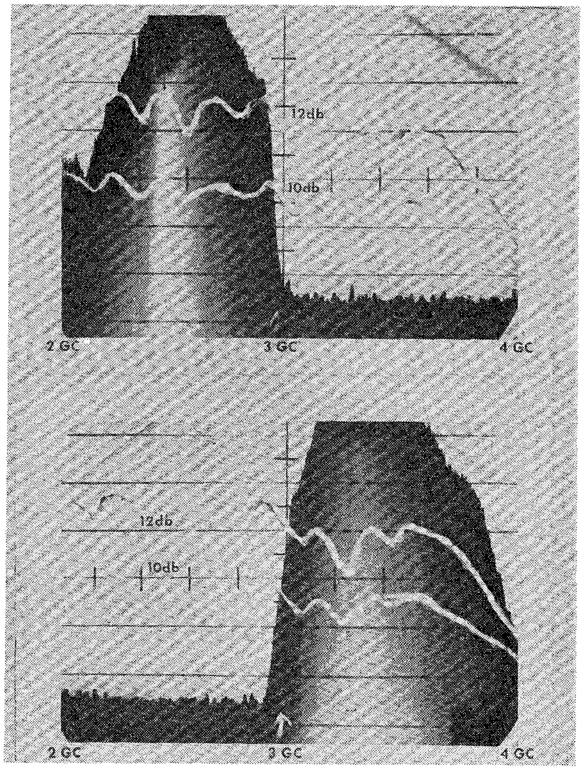

Fig. 9—Measured gain response of half-octave amplifiers.

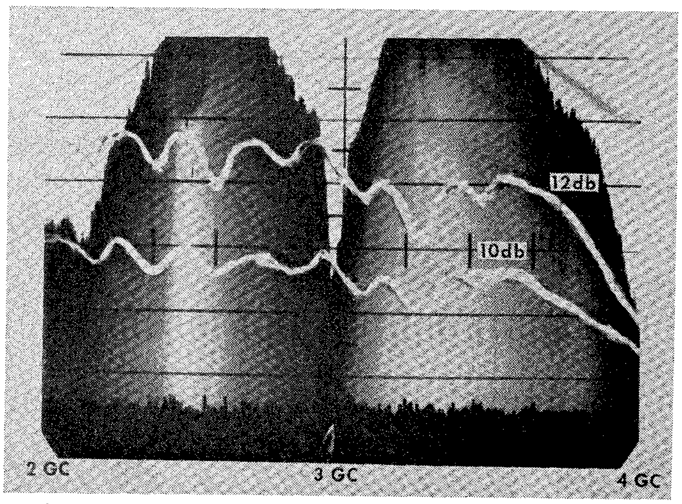

Fig. 10—Superimposed gain responses of half-octave amplifiers.

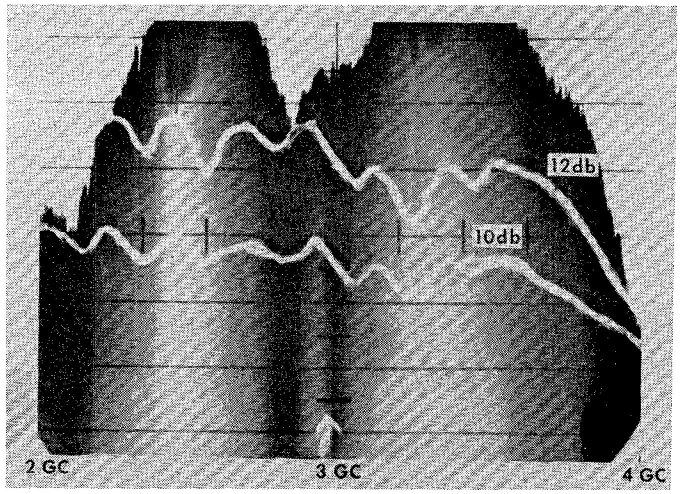

Fig. 11—Measured gain response of octave amplifier.

Multiplexing Two Amplifiers

An octave amplifier was formed by combining the 2- to 3-Gc amplifier with a 3- to 4-Gc amplifier designed in a similar manner. The technique used to avoid crossover loss has been described by Kelch.[6] As implied by the reference title, the technique is applicable to multioctave designs. Fig. 8 is a photograph of the octave amplifier. The low-pass filters have a 3-Gc cutoff frequency. Incoming signals in the 3- to 4-Gc band are reflected by the filters into the circulator. Similarly, the circulator output is reflected by the filters to the amplifier output. Signals in the 2- to 3-Gc band arrive at the output via the other path—through the filters and the 2- to 3-Gc amplifier.

In the crossover region, the signal splits into both paths, recombining at the output port. Fig. 9 is an oscilloscope photograph showing the gain response when the bias voltage of each amplifier is disconnected in turn. The lower reference line corresponds to 10-db gain; the upper, to 12-db gain. Fig. 10 shows the two pictures in Fig. 9 superimposed, while Fig. 11 shows the octave amplifier response with both diodes biased. The crossover signals are in phase when they recombine because the path length of the higher frequency amplifier was adjusted by the phase correctors appearing in Fig. 8.

Conclusions

The design of half-octave tunnel-diode amplifiers based on a Chebyshev low-pass prototype has been described. The amplifiers were combined in a multiplex circuit to provide an octave bandwidth amplifier. An alternate design procedure was developed after the units were built. This method is simpler and more accurate. When circulators with a broader bandwidth become available, the described procedure may be used to design amplifiers of matching bandwidth. The multi-

[6] R. D. Kelch, "Multiple-octave low-noise microwave preamplifier," *Microwave J.*, pp. 75–82; May, 1963.

plexing technique used to form the octave amplifier is adaptable to multi-octave designs. The measured gain responses given in Fig. 9 do not follow the theoretical curves of Figs. 3 and 4, because of various simplifications in the analysis. For instance, the series inductance of the diode was neglected, the load was assumed to be 50 Ω and discontinuities in the structure were neglected. At higher frequencies, these simplifications will result in a greater discrepancy between computed and measured performances.

Appendix

Design Equations for Double Tuned Tunnel-Diode Amplifier

The elements B_1 and X_2 in Fig. 1 are evaluated by making the input reactance of the circuit vanish at band edges while the input resistance is determined by the equation for power gain,

$$G = \frac{(R_g - R)^2}{(R_g + R)^2}.$$

At center frequency, the elements are assumed to be resonant by symmetry considerations and R_D is then determined by center frequency gain. Let us assume 11-db gain, so that $R_D = -89.2\ \Omega$.

The shunt element is evaluated by setting the real part of the impedance equal to $-28\ \Omega$ at both band edges, corresponding to 11-db gain.

Since the series reactance cannot contribute to the real part of the circuit impedance, only the shunt element need be calculated. It consists of a shorted length of transmission line shunted across the diode capacitance.

Real part of Z = Real part of

$$\frac{1}{j\omega C - jY_{01}\cot\theta_1 - \dfrac{1}{89.2}} = -28,$$

$$\frac{1}{\left(\dfrac{1}{89.2}\right)^2 + (\omega C - Y_{01}\cot\theta_1)^2} = -28,$$

$$\omega C - Y_{01}\cot\theta_1 = \sqrt{\frac{1}{28 \times 89.2} - \left(\frac{1}{89.2}\right)^2}$$

$$= \pm 0.01656.$$

Using subscript 2 for upper band edge and subscript 1 for lower band edge,

$$Y_{01}\cot\theta_{12} = \omega_2 C \pm 0.01656$$

and

$$Y_{01}\cot\theta_{11} = \omega_1 C \pm 0.01656.$$

Dividing to eliminate Y_{01} and using f in Gc and C in pf for convenience,

$$\frac{\tan\theta_{11}}{\tan\theta_{12}} = \frac{f_2 C - 2.64}{f_1 C + 2.64}.$$

The signs were chosen to make this quantity less than unity in order to get solutions in the first quadrant.

Since

$$\theta = \frac{\omega l}{c},$$

we may write

$$\theta_{11} = \frac{f_1}{f_2}\theta_{12}$$

and

$$\frac{\tan\dfrac{f_1}{f_2}\theta_{12}}{\tan\theta_{12}} = \frac{f_2 C - 2.64}{f_1 C + 2.64}.$$

The other equation for the shunt element is a rearrangement of one of the pair above with proper constants for expressing f in Gc and C in pf.

$$Y_{01} = (0.00628C - 0.01656)\tan\theta_{12}.$$

The equations for the series element are derived from the condition that the imaginary part of the circuit impedance vanishes at both band edges.

$$Z_{02}\tan\theta_2 + \frac{Y_{01}\cot\theta_1 - \omega C}{\left(\dfrac{1}{89.2}\right)^2 + (\omega C - Y_{01}\cot\theta_1)^2} = 0$$

and $Y_{01}\cot\theta_1 - \omega C = \pm 0.01656$, with the positive sign for the upper frequency and the negative sign for the lower frequency.

Combining these leads to the following pair of equations:

$$Z_{02}\tan\theta_{21} = -41.4,$$

$$Z_{02}\tan\theta_{22} = +41.4.$$

Dividing to eliminate Z_{02},

$$\frac{\tan\dfrac{f_1}{f_2}\theta_{22}}{\tan\theta_{22}} = -1.$$

Let us assume that the series resonator is a half wavelength at center frequency to satisfy the symmetry conditions we have assumed. Then θ_{21} must be less than 180° by the same amount, θ, that θ_{22} is more than 180°.

$$\theta_{21} = 180° - \theta,$$

$$\theta_{22} = 180° + \theta.$$

But

$$\theta_{21} = \frac{f_1}{f_2}\,\theta_{22}.$$

Therefore

$$180° - \theta = \frac{f_1}{f_2}\,(180° + \theta)$$

or

$$\theta = \frac{f_2 - f_1}{f_2 + f_1}\,180°.$$

This relationship is a function only of the ratio of band edge frequencies. The characteristic impedance of the series element is

$$Z_{02} = \frac{41.4}{\tan\theta_{22}} = \frac{41.4}{\tan\theta}\,.$$

This quantity also is a function of the ratio of band-edge frequencies, independent of center frequency. The constant is determined by the choice of gain at center frequency and at band edges. For example, octave bandwidth at 11-db gain corresponds to 23.9 Ω.

Acknowledgment

The low-pass filters in the multiplexer were designed by H. Yee. The experimental work resulting in the gain response curves of Figs. 9, 10 and 11 was performed by A. Camps. The no-prototype design method resulted from many discussions with R. D. Hall.

Microwave Breakdown Near a Hot Surface*

M. GILDEN†, MEMBER, IEEE, AND J. PERGOLA†, MEMBER, IEEE

Summary—Microwave breakdown near a hot surface in a waveguide system was studied to determine its dependence upon the thickness of the adjacent film of hot gas and its associated temperatures. The effect of the variation of the film thickness with the flow rate of the bulk of the gas was of particular interest. To carry out the theoretical analysis, a more general breakdown equation was derived to account for the temperature gradients. Experimental results supporting the theory also are presented.

The study shows that, although the breakdown threshold of a waveguide system is lowered by the presence of a hot surface, a sufficiently rapid flow of the bulk gas tends to restore the threshold as a result of the reduction in the thickness of the film of hot gas. This effect occurs in addition to that reduction resulting from cooling the surface.

Introduction

One of the problems associated with maintaining the high power capabilities of microwave transmission lines and components is the reduction of the breakdown threshold resulting from localized heating. Breakdown may be initiated in the region of heated gas at field strengths much below that required for breakdown of the main volume, since the field strength at which ionization begins to occur is inversely proportional to the absolute value of the temperature of the gas in a constant pressure system. In the presence of localized heating, temperature gradients are estab lished in the gas adjacent to the hot surface which in turn leads to nonequilibrium conditions characterized by convection currents. The motivation for this study was the realization that the layer of hot gas may consist of only a thin film whose thickness can be controlled by the velocity of the gas flow across the surface.[1] Thus, breakdown in the film can also be controlled.

Rapid gas flow may also contribute to an increased rate of electron loss from the region of ionization with the consequence that the breakdown threshold is raised. This effect has been reported at low pressures[2] (10 mm Hg); however, at the atmospheric pressure used in the current work, the effect was absent.

The problem of microwave breakdown under uniform conditions[3] or nonuniform conditions of electric field have been investigated over the last few years.[4,5] Practical considerations of breakdown problems in

* Received May 20, 1963; revised manuscript received August 20, 1963. This work was sponsored by Rome Air Development Center, Rome, N. Y., Contract No. AF30(602)-2545.

† Microwave Associates, Inc., Burlington, Mass.

[1] W. H. Giedt, "Principles of Engineering Heat Transfer," D van Nostrand Co., Inc., New York, N. Y.; 1957.

[2] J. G. Skinner and J. J. Brady, "Effect of gas flow on the microwave dielectric breakdown of oxygen," *J. Appl. Phys.*, vol. 34, pp. 975–978; April, 1963.

[3] L. Gould and L. W. Roberts, "Breakdown of air at microwave frequencies," *J. Appl. Phys.*, vol. 27, pp. 1162–1170; October, 1956.

[4] M. A. Herlin and S. C. Brown, "Electrical breakdown of a gas between coaxial cylinders," *Phys. Rev.*, vol. 74, pp. 910–913; October, 1948.

[5] P. M. Platzman and E. Huber-Solt, "Microwave breakdown in nonuniform electric fields," *Phys. Rev.*, vol. 119, pp. 1143–1149; August, 1960.