Some Fundamental Design Principles for the Development of Precision Coaxial Standards and Components

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Abstract—Significant advances in the performance of precision coaxial standards and components have resulted from rigid adherence to three basic design principles: 1) incremental constancy of characteristic impedance, 2) coplanar compensation of discontinuities, and 3) control of mechanical tolerance sensitivity. The advances include dielectric supports with extremely small interface discontinuities, contacting members that are insensitive to mechanical tolerances, calculable airline impedance standards of very high absolute accuracy, nearly reflectionless terminations based on a cylindrical metal-film resistor, and adaptors between line sizes based on smooth (rather than stepped) diametral transitions. Extensions of the design principles have resulted in the introduction of closely controlled reflections provided by a novel impedance-matching tuner, broadband calibrated mismatches, and resistance-standard terminations that retain their nominal dc values to very high frequencies.

DESIGN PRINCIPLES

Introduction

RECENTLY-developed precision coaxial connectors [1], measuring equipments [2], [3], standards and components are characterized by extremely low residual reflections from dc to the cutoff frequency of the line size in use. To achieve such performance, it is necessary to employ three principles of design that theoretically eliminate impedance discontinuities on an incremental basis; that is, each incremental section of transmission line is adjusted to introduce no reflection. Conventional designs employing compensation sections that are not short compared with a wavelength result in narrow-band, less-than-optimum performance.

Design Principle 1

Maintain a constant characteristic impedance on an incremental basis whenever possible. In many coaxial devices, a constant characteristic impedance is not maintained at all transverse cross sections. Sections of line with characteristic impedances above and below the nominal are often employed to compensate for steps in conductor diameters, slots in conductors, or gaps (very short sections of overcut outer conductors or undercut inner conductors). Such designs cannot be employed in broadband precision devices. Steps, slots, gaps, and large variations in diameters of the inner or outer conductor introduce impedance discontinuities, which, in turn, limit bandwidth and performance. The characteristic impedance at each cross section throughout a device (i.e., the incremental characteristic impedance) must be maintained as closely as possible to the nominal value (e.g., 50 ohms) to achieve the optimum broadband performance. For example, the air-line standards described in this paper are simply sections of constant-characteristic-impedance coaxial lines. The SWR introduced by these sections is less than 0.05 percent[1] from dc to 9 GHz.

Design Principle 2

Introduce an individual coplanar compensation for each unavoidable impedance discontinuity. Impedance discontinuities cannot always be avoided. For example, a constant characteristic impedance cannot be maintained both inside and outside a dielectric support for a coaxial line without stepping one or both conductor diameters and consequently introducing impedance discontinuities. For best performance in such instances, it is necessary, first, to minimize the uncompensated discontinuity and, second, to introduce an individual coplanar compensation for the disturbance that remains.

Coplanar compensation is that compensation which appears electrically to be distributed over the same region as the original disturbance; it thus results in the broadest frequency bandwidth of performance. The common practice of changing the characteristic impedance of a long section of line to compensate for a lumped discontinuity restricts bandwidth and should be avoided.

Design Principle 3

Minimize dependence of electrical performance on mechanical tolerances. Tolerances on conductor dimensions cannot be avoided in coaxial devices; often, however, several mechanical tolerances contribute to a single conductor diametral tolerance. For example, the net tolerance on the diameter of the inner conductor of a UG jack connector (such as a Type N jack connector) depends on the tolerances on three diameters: the outer diameter of the slotted fingers, the inner diameter of the slotted fingers, and the outer diameter of the contact pin of the mating-plug inner conductor.

It is important, therefore, to allow only *one* mechanical tolerance to affect an electrically important dimen-

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[1] SWR in percent is given by $(SWR-1)\times 100$.

sion, and this tolerance should apply to a mechanical dimension *not subject to wear.* For example, in the inner-conductor contact arrangement described in this paper, the contact is inside the conductor so that the inner-conductor diameter does depend on just one mechanical tolerance and the electrical performance is independent of contact wear.

Precision Coaxial Transmission Lines

Air Lines

Sections of constant-impedance air-dielectric coaxial line are essential to almost all coaxial devices and are the bases for calculable impedance standards at microwave frequencies. They are nearly reflectionless and represent the ultimate in adherence to the design principles: A constant characteristic impedance is maintained throughout (Design Principle 1). Since there are no impedance discontinuities, no compensation is required (Design Principle 2). Electrical performance depends primarily on just the two wear-independent conductor diameters (Design Principle 3).

The effect of conductor diameter errors on characteristic impedance for a 50-ohm coaxial line is given by

$$\Delta Z_0 = 0.120 \frac{\Delta D - 2.3\Delta d}{D}, \text{ percent} \qquad (1)$$

where:

ΔZ_0 is the characteristic impedance error in percent,
D is the outer-conductor diameter in inches,
ΔD is the error in the diameter of the outer conductor in mils,
Δd is the error in the diameter of the inner conductor in mils.

Therefore, a characteristic-impedance tolerance of 0.05 percent in a $\frac{9}{16}$-inch 50-ohm line implies an outer-conductor diameter tolerance of 120 microinches and an inner-conductor diameter tolerance of 50 microinches.

Measurement of Conductor Diameters

There are many practical difficulties in determining the effective diameter of a conductor to an accuracy of 5 or 10 microinches. Some critical factors are the ellipticity and triangularity of the conductor cross section, the mechanical standard of diameter, errors in the measuring system, and the measurement environment.

Ellipticity and triangularity of conductor cross section are second-order electrical effects and are important only as they complicate measurements of the average conductor diameter. Ellipticity can be averaged out by three-port air gauges [4]; triangularity can be averaged out by two-port air gauges. Capacitance gauges [4], [5] can also be used to measure the average diameters of conductors.

Mechanical standard plugs and rings are available with diameter accuracies of 5 microinches, and conventional air-gauge measuring systems are accurate to within 10 microinches. Temperature, humidity, and pressure effects can be eliminated by means of a controlled environment.

Effects of Surface Finish

Surface finishes below about 15 microinches tend to average out electrically; however, some observations indicate a dependence of characteristic impedance on surface finish. The characteristic impedance of a $\frac{9}{16}$-inch reference air line was found to differ more than 0.1 percent at 7 GHz when an inner conductor with a surface of 75 microinches was substituted for one with a 5-microinch finish. The direct change in conductor diameter corresponding to such a difference in characteristic impedance is approximately 200 microinches, yet the actual difference in inner-conductor average diameters was less than 20 microinches! While no theoretical explanation of this effect has yet been found, it is measurable and should be noted.

Effects of Conductor Plating

Conductor surfaces are plated to decrease resistivity but often with disappointing results [6]. Conductor resistivity is important because it enters into the relations for characteristic impedance and velocity of propagation as well as into the loss relations [7]. The effects of conductor resistivity above 500 MHz are second-order on characteristic impedance and velocity of propagation and are usually negelected; however, accurate knowledge of the resistivity, particularly as a function of frequency, permits calculation of these effects as well as the loss.

The conductor resistivity that can be achieved with silver plating varies widely with the plating technique as can be seen in Fig. 1. Conductors silver-plated in a bath containing commercial brighteners exhibit resistivities of the order of brass. Conductors silver-plated in baths without commercial brighteners yield much lower resistivities. Periodically reversing current in the plating process (*P-R* process) [8] lowers effective resistivity further at the higher frequencies. The explantion for this improvement appears to be that dc plating is more porous than *P-R* plating. Dc-plated parts that are cold-worked to compact the plating do not exhibit the marked increase in resistivity at higher frequencies. Cold-working is not required with *P-R* plated parts. The measured resistivities of polished brass and Consil® (Ag-Mg-Ni alloy) inner conductors are independent of frequency, which agrees with theory and checks the experiment.

A number of conclusions can be drawn from the above:

1) Commercial brighteners added to silver-plating baths may increase the resistivity so much that

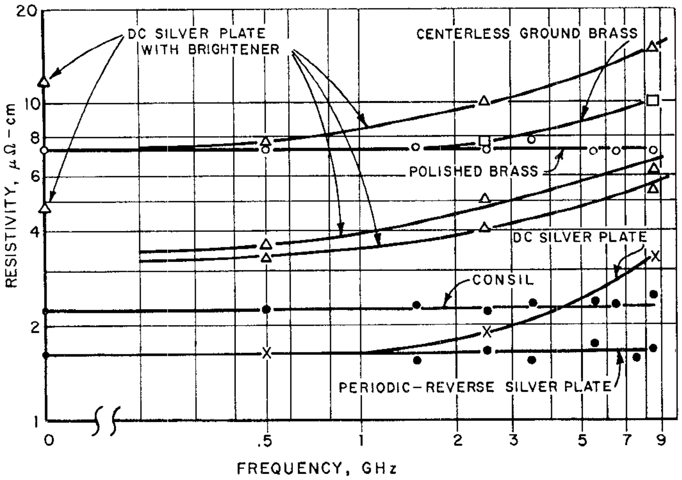

Fig. 1. Measured data on conductor resistivity vs. frequency.

the conductor resistivity is actually worse than that of brass.

2) Dc-plated parts have a resistivity that rises at high frequencies.
3) The periodic-reverse plating process yields high-frequency resistivity equal to that of pure silver.
4) If conductors are to be cold-worked after plating, dc and *P-R* plating processes yield similar results. (Note that cold-working of plated parts where the plating contains brighteners does *not* lower the high-frequency resistivity characteristic. The plating apparently does not compact.)

The reference standard air lines described in this paper are silver-plated by the *P-R* process.

Precision Connectors

An Ideal Connection of Two Coaxial Lines

An ideal (and hypothetical) connection of identical precision coaxial lines is shown in Fig. 2(a). Butt joints are achieved simultaneously at both the outer- and inner-conductor junctions. The aligning ring makes the outer conductors concentric across the junction. The dielectric supports, which are of unity dielectric constant and reflectionless, support and center the inner conductors. This connection is unattainable in practice, of course, but it does represent a bench-mark for the evaluation of practical precision connector designs.

A Practical Connection

A practical, but nearly ideal, connection is shown in Fig. 2(b). Here the dielectric supports (with a dielectric constant of about 2) both radially and axially support the inner conductors. A butt joint is achieved at the outer-conductor junction, but, owing to axial mechanical tolerances, a butt joint cannot be achieved at the same time at the inner-conductor junction. Therefore, to prevent damage or motion of the inner conductor with respect to the outer conductor, the inner conductors are slightly recessed behind their respective outer conductors, and a flexible contact bridges the small resulting inner-conductor gap. An aligning ring centers the outer condutors.

The connection described deviates from the cross section of a precision coaxial line or from the ideal connection in only two areas: the region of the dielectric supports and the region of the inner-conductor contact.

Dielectric Support Design

For optimum performance over the widest possible frequency range, the characteristic impedance inside a support must be the same as that of the associated air-dielectric line (Design Principle 1). To achieve this condition, it is necessary to undercut the inner conductor, overcut the outer conductor, or do both as illustrated in Fig. 2(b). The undercutting and overcutting introduce unavoidable discontinuity capacitances at the support faces; however, these discontinuities can be minimized by the proper undercut-overcut combination as shown in Fig. 3.

Here the "total" curve represents the net discontinuity capacitance at a support face as a function of overcut in the outer conductor for a 50-ohm Teflon support. (The characteristic impedance in the dielectric region is constant under all conditions shown as required by Design Principle 1.) The total discontinuity capacitance at a support face is the sum of the individual inner- and outer-conductor step discontinuity capacitances [9]. The total discontinuity is minimum when the outer conductor overcut is approximately 20 percent of a fully overcut design. Moreover, the performance of this support is then least sensitive to mechanical tolerances (Design Principle 3).

Coplanar compensation of the remaining interface disturbances (Design Principle 2) is well-approximated by removal of material from the faces of the support [10] [Fig. 4(a)]. For a short support, some interaction is encountered in effecting face compensations and maintaining the constant characteristic impedance in the dielectric region. However, convergence on the optimum design by adjustment of face-compensation and characteristic-impedance parameters is relatively fast.

Figure 4(b) is a convenient presentation of impedance data for such a support. The ordinate is the imaginary component of the normalized impedance, referred to the center of the support and expressed in percent. The abscissa is frequency in GHz. If the support is symmetrical about its center and the reflections involved are small, then only imaginary components should be present when so referenced, and the ordinate is simply SWR in percent. (The presence of real components indicates support asymmetry or measurement error and can be treated separately.) The crosses on the plot are based on measured data taken for the Teflon support

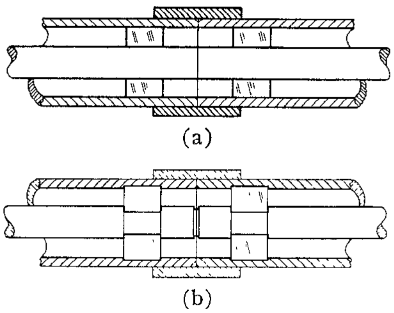

Fig. 2. Coaxial transmission line junctions: (a) ideal, (b) practical.

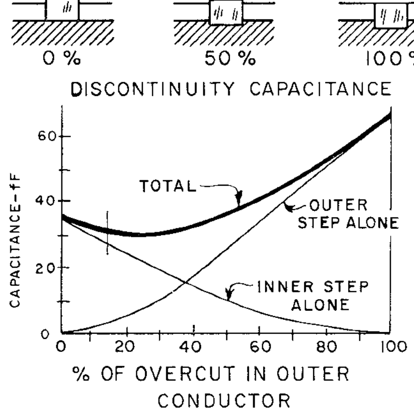

Fig. 3. Inner-conductor and outer-conductor step discontinuities at the interface of a 50-ohm Teflon Support for a 9/16-inch, air dielectric coaxial transmission line.

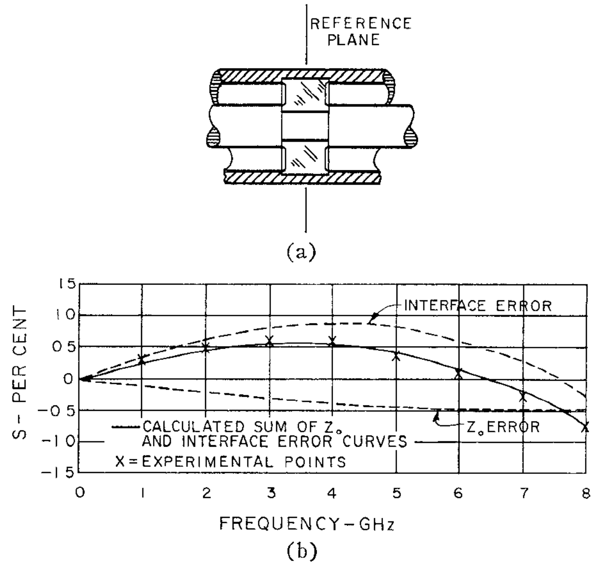

Fig. 4. Dielectric support: (a) cross section showing face compensation, (b) analysis of SWR performance.

described earlier, at a stage of development when both the characteristic-impedance and interface compensation errors were still appreciable. The amounts of each are determined by the proper choice of amplitude and phase (plus or minus) of two well-defined curves (shown dashed) that, superimposed, match the measured data.

The SWR caused by a small error in characteristic impedance is given by:

$$S = 2\Delta z \sin \frac{2\pi f}{f_0} \tag{2}$$

where:

S is the SWR in percent,
Δz is the characteristic-impedance error in percent (positive if higher than nominal, negative if lower than nominal),
f is the frequency in GHz,
f_0 is the frequency in GHz at which the electrical length of the support is one wavelength.

The SWR caused by a small error in interface compensation is given by:

$$S = 2f\Delta x \cos \frac{2\pi f}{f_0} \tag{3}$$

where S, f, and f_0 are as before and Δx is the SWR of a single interface in percent at 1.0 GHz (positive if the interface disturbance is inductive and negative if it is capacitive).

For the Teflon support, f_0 is 30 GHz; the support is one cm long electrically. For this example, the proper choice is $\Delta z = -0.24$ percent and $\Delta x = +0.17$ percent for which the dashed curves are obtained. Combination of the dashed curves results in the solid curve shown, which agrees closely with the measured data. The characteristic impedance and the interface compensation can now be corrected, new measurements taken, and a new analysis made in a similar manner. The design-center support developed with these design principles introduces an SWR of less than 1.001 from dc through 9 GHz.

Dielectric Support Control

Air gaps between dielectric and metal surfaces have a large, unpredictable effect on electrical performance. Air gaps are eliminated when the support is a press fit in all directions in the associated metal parts; moreover, the degree to which mechanical tolerance affects electrical performance (Design Principle 3) is then minimized. Final support diameters are then determined by the metal dimensions.

For a Teflon support less than a quarter-wavelength long, the *weight* of the support has a first-order effect on performance. The effective dielectric constant of the compressed support and the indentations of the support faces have only second-order effects *when the weight is maintained constant.*

As an illustration, consider the reflection introduced at a frequency where the support is less than a quarter-wavelength long. Suppose the outer diameter of the support is above nominal and the inner diameter is below nominal; then, when the support is compressed between design-center metal parts, a below-nominal characteristic impedance and a corresponding capacitive effect is introduced. Since the support weight has been kept constant, an excess of material must have been removed from the support faces, resulting in overcompensation of the interface discontinuities. This inductive effect tends to cancel out the capacitive effect caused by the diameter errors. The converse situation is also self-compensating. Therefore, the support weight should be considered in design and subsequently maintained as close as possible to nominal.

Figure 5 illustrates the effect of weight control in a somewhat different way. Supports with nominal diameters were fabricated from three types of Teflon, one of nominal dielectric constant and the other two at the extremes available. The weight of all supports was maintained constant as described earlier. All three supports were measured with the same metal parts under the same conditions. Taking the data for the nominal-value support as a reference, the data on the extreme-value supports are plotted in Fig. 5 as solid curves. The extreme dielectric constants were +0.45 percent and −0.48 percent from nominal. The dashed curves represent the corresponding calculated error curves without weight control.

Note that the weight control works exceptionally well up to 5 GHz (where the support is approximately $\frac{1}{6}$ wavelength). Weight control theoretically has no effect where the support is a quarter-wavelength long and has a slightly detrimental effect when the support is more than a quarter-wavelength long. However, the control of lower-frequency performance is well worth the small sacrifice in higher-frequency performance. The Teflon dielectric constant can be controlled in practice to about 0.25 percent.

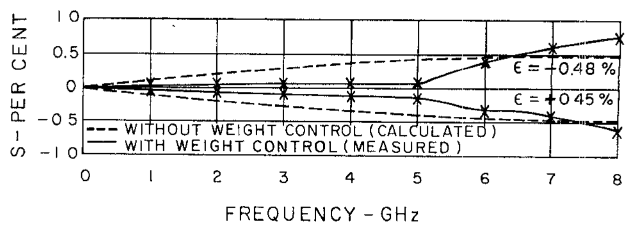

Fig. 5. Illustration of the effect of weight control on support performance.

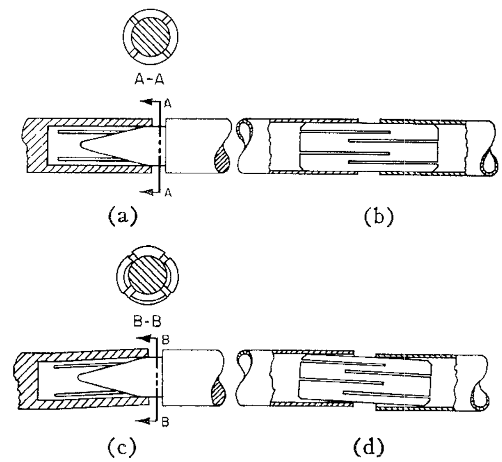

Fig. 6. Inner-conductor contact arrangements: (a) bayonet, (b) bullet, (c) bayonet when inner conductors are eccentric to each other, (d) bullet when inner conductors are eccentric to each other.

Contact Design

Two commonly used inner-conductor contact arrangements are shown in Fig. 6: (a) the overlapping or bayonet contact arrangement and (b) the bullet. With the bayonet arrangement, allowance for axial tolerances is provided by the gap between the end of the spring fingers and the step on the bayonet. This gap introduces an appreciable reflection. Allowance for eccentricity is provided by the ductility of the spring fingers; however, when the fingers are distorted as shown in (c), the contact becomes unreliable. With the bullet arrangement, allowance for axial tolerances is also provided by a gap between inner-conductor tips. Allowance for eccentricity is very limited (particularly when the inner-conductor tips are close together) and as shown in (d) the contact becomes unreliable.

A butt-joint inner-contact arrangement that solves the problems outlined and is nearly reflectionless is illustrated in Fig. 7. In this arrangement, there are no male or female parts. All connectors are identical. Low reflection is achieved because the pattern of current flow deviates only slightly from that of the ideal connector of Fig. 2(a).

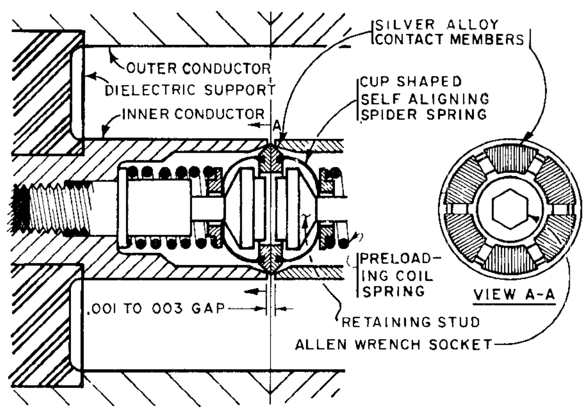

Fig. 7. Butt-joint inner-conductor contact arrangement.

This unique contact arrangement provides a sliding or wiping action in all contact areas. This wiping, self-cleaning action is the action that ensures reliable connection but does *not* involve any of the critical electrical diameters.

With this contact arrangement, a constant characteristic impedance is maintained (Design Principle 1) except in the region of the very small gap between the inner conductors. The gap is the only impedance discontinuity and it is only 1 to 3 mils by 10 mils. Compensation (Design Principle 2) is not required because the gap disturbance is negligible. Axial tolerances and contact-diameter tolerances introduce only small second-order effects on the electrical performance (Design

Principle 3). A contact arrangement of this type for the inner conductor of the $\frac{9}{16}$-inch line size introduces an SWR calculated to be approximately 1.001 at 9 GHz.

This butt-joint contact arrangement also offers many other advantages. For example, only one contact is required for good electrical connection, since the spring contact will mate just as well with a flat surface. A good contact is also achieved when the two mating inner conductors are slightly eccentric and, equally important, no moments or increased forces are caused by the eccentricity.

The dielectric support and the contact described before are the essential elements of a precision coaxial connector which meets or exceeds the requirements of the IEEE Recommended Practices on General Precision Connectors for the $\frac{9}{16}$-inch (14-mm) line size [11]–[13].

Standards

Reference Standard Air Lines

A section of precision coaxial transmission line is the most accurate standard of microwave impedance available [7], [14], [15]. Such a section (with a second-order slot correction) is the built-in impedance standard of coaxial slotted lines. It is the impedance standard in sliding loads and sliding short circuits, and it is also the reference standard in quarter-wave, half-wave, and other substitution techniques.

Practical application of this standard has been limited until recently since, in order to use the air line, it is necessary to equip it with connectors, which, however precise, reduce the usable air-line accuracy. Required is a means of using the air line without employing connectors. One means of accomplishing this employs the specially designed reference air line shown in Fig. 8.

Here the butt-joint outer-conductor connection is provided with a suitable aligning ring. Retractable spring-loaded pins support the inner conductor by centering in the mating connectors. The air line "connectors" are thus not really connectors at all, but simply coupling and support mechanisms. As such, they contribute nothing to the residual VSWR of the standard. Connectors without dielectric supports, such as these, are called Laboratory Precision Connectors [11], [12]. (Precision connectors with dielectric supports are called General Precision Connectors [11], [12].)

Reference air lines, if properly constructed, are constant-impedance precision coaxial lines and, thus, absolute, or calculable, impedance standards. Figure 9 shows the residual SWR of such a reference air line for the $\frac{9}{16}$-inch line size. Note that it is less than 1.0005 over most of the frequency range.

The uses of reference air lines as impedance standards in a number of substitution techniques with one- and two-port networks are described by Sanderson [16], [17], Zorzy [18], and Harris [7]. The residual reflections of measuring instruments such as slotted lines, bridges,

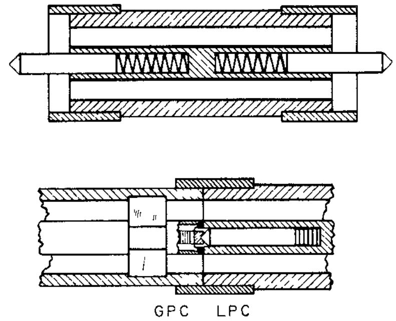

Fig. 8. Top—Schematic cross section of a reference impedance air line; Bottom—Junction of precision coaxial connector and reference impedance air line.

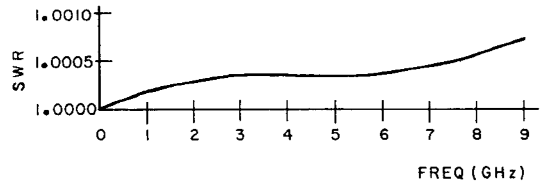

Fig. 9. Measured SWR introduced by a reference impedance air line developed for the 9/16-inch line size.

directional couplers, and hybrids, the residual reflections of connectors, and the residual reflections of terminations are among those that can be separated and measured to accuracies of better than 0.1 percent (equivalent SWR 1.001) by these standards and substitution techniques.

Reference air lines are also precision time-delay, reactance, and capacitance standards. Constant characteristic impedance and close control on axial tolerances achieved with practical $\frac{9}{16}$-inch reference air lines provide time delays up to 1000 picoseconds ± 0.1 picosecond or better. Reactance standards comprise a reference air line and a short circuit whose reactance is calculable from the frequency and the characteristic impedance and electrical length of the reference air line. Low-frequency capacitance standards are also calculable from the reference air-line dimensions and are limited in accuracy only by the mechanical tolerances on the air-line dimensions and the eccentricity of the air-line conductors. Agreement between calculated and measured capacitance at 1 KHz for a 10-picofarad (15 cm) $\frac{9}{16}$-inch air-line standard is better than 0.005 pF. Connection repeatability of capacitance is better than 0.001 pF.

Matched Terminations

The theoretical design of a matched resistive-film termination that introduces no local impedance disturbances has been presented by Harris [19]. The inner conductor is a cylindrical film resistor; the outer conductor is a tractrix. A reflectionless connection into such a matched termination from a section of precision coaxial line (characterized by a planar field configuration) is complicated by the spherical field at the termination input.

A section of constant-impedance transmission line (Design Principle 1) that smoothly bends the field from a planar to a spherical configuration and is essentially free of impedance discontinuities is shown in Fig. 10(a). The outer-conductor contour is an arc of a parabola (or circle) tangent to the outer conductor of the section of precision line at one end and tangent to the outer conductor tractrix of the termination at the other end. Assuming a parabolic curve, the outer conductor diameter (b) is given at any point (l) (relative to the junction with the planar field) by

$$b = b_0 - \frac{\tan^2 \theta_1}{4(b_0 - b_1)} l^2 \tag{4}$$

where the various dimensions are as labeled in Fig. 10. The distance l_1 is fixed, since both the diameter and slope are specified at point 1 (the junction with the tractrix), and is given by

$$l_1 = \frac{2(b_0 - b_1)}{\tan \theta_1}. \tag{5}$$

The radius associated with the field configuration is thus decreased from infinity (the planar field) to that of the spherical field at the termination input.

Incrementally, the transition section is a section of a conical transmission line with a common apex for both conductors, whose characteristic impedance is defined in terms of the slopes of the two conductors [19], as

$$Z_0 = 59.96 \ln \frac{\tan \theta_o/2}{\tan \theta_i/2} \tag{6}$$

where θ_o and θ_i are as shown in Fig. 10. Therefore, the inner-conductor diameter a corresponding to the outer-conductor diameter b (a and b being located on the spherical wavefront) can be determined through the use of (6) as

$$a = \frac{\sin \theta_i}{\sin \theta_o} b. \tag{7}$$

The offset Δl between the longitudinal positions of a and b is given by

$$\Delta l = \frac{a}{2 \tan \theta_i} - \frac{b}{2 \tan \theta_o} \tag{8}$$

based on the condition of a common apex for the incremental outer- and innner-conductor tangents.

Thus, the inner-conductor diameter is defined uniquely at each point in terms of the outer-conductor diameter, the outer-conductor slope, and the desired characteristic impedance. There are no impedance discontinuities at the junction of the conventional air line and the transition such as there would be with a straight conical line. Also, the slower the rate of change of the conductor slope in the transition region, the more nearly does the line approximate sections of conical transmission line. In a specific 50-ohm termination, a

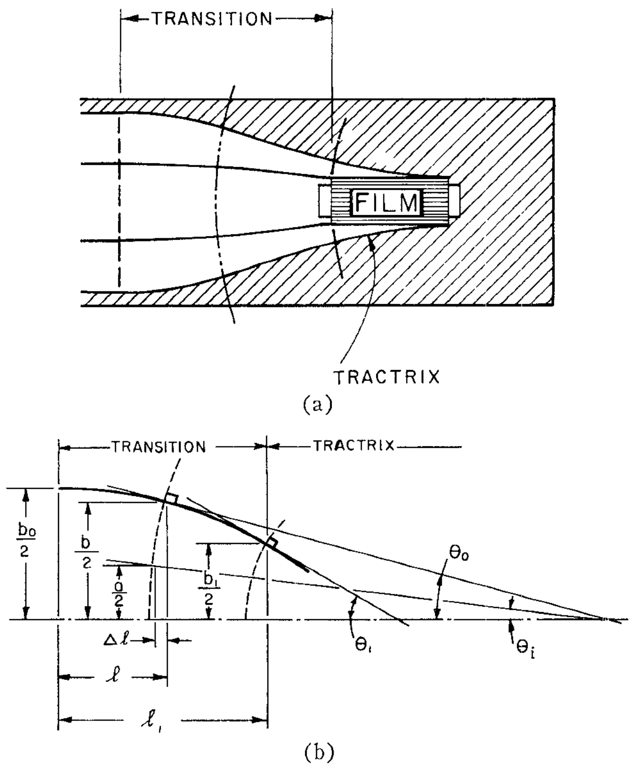

Fig. 10. Matched termination with a smooth transition from the planar field configuration to the spherical field configuration: (a) cross-section view, (b) geometry of transition.

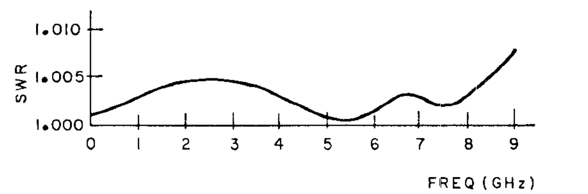

Fig. 11. Optimum performance achieved for a matched 50-ohm termination in the 9/16-inch line size.

change in slope from 0° at the $\frac{9}{16}$-inch diameter to 15° at a diameter of $\frac{5}{16}$ inch (at the tractrix interface) was accomplished with no apparent discontinuities in the transition section.

Practical limitations in the fabrication of the resistive elements are a problem. The diameter and surface condition of the substrate on which the resistive film is deposited, the length, uniformity and longitudinal location of the film on the substrate, and the means of mechanical connection between the conductors and the film are critical considerations that have led to small deviations from the ideal tractrix and, at present, limit the performance of this type of matched termination.

Figure 11 shows the optimum performance achieved for such a 50-ohm termination equipped with a $\frac{9}{16}$-inch precision connector. The VSWR is less than 1.005 from dc to 8 GHz.

Calibrated Mismatches

Broadband calibrated mismatches result if the 50-ohm resistive element used in the matched termination

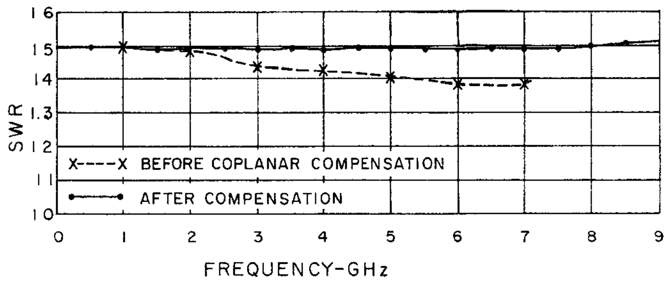

Fig. 12. Measured performance of calibrated mismatch before and after coplanar compensation.

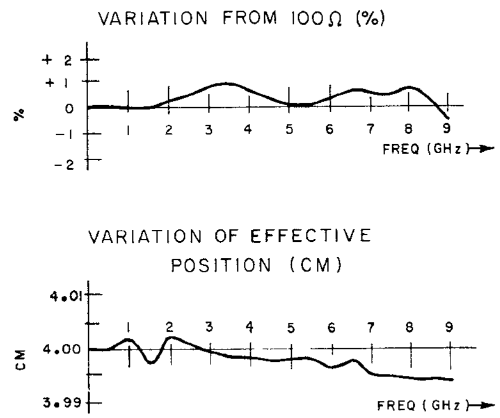

Fig. 13. Measured characteristics of a 100-ohm resistive termination for a 50-ohm 9/16-inch coaxial line.

is replaced by resistive elements of other than 50 ohms.

The specific performance requirement on a calibrated mismatch is that the reflection introduced into an otherwise matched system remains constant over a broad frequency range. Such a device is extremely useful for direct RF system checks on slotted lines, bridges, reflectometers, and other impedance-measuring systems.

Outer-conductor tractrices based on resistances other than 50 ohms are inconvenient because of the resulting large capacitive discontinuity at the junction of a tractrix (at its characteristic-impedance level) and the 50-ohm air line or transition. On the other hand, the 50-ohm tractrix is not adequate without modification. Therefore, coplanar compensation (Design Principle 2) was introduced at the air-line resistor junction and along the (50-ohm) outer-conductor tractrix using experimental data to locate and measure impedance discontinuities.

Figure 12 shows the performance before and after coplanar compensation of a calibrated mismatch that introduces an SWR of 1.5. Note that the variation in SWR of the compensated unit is less than 0.02 throughout the 0–9-GHz frequency range.

Resistance-Standard Terminations

The performance requirements on a resistance standard (other than a matched termination) are that its effective resistance remain equal to its dc resistance and that its effective position remain constant over a broad frequency range. Such standards are particularly useful in the direct RF calibration of bridges and complex reflection-coefficient measuring systems.

Broadband resistive standards at the 100- and 200-ohm levels have been achieved in a manner similar to that used in obtaining broadband calibrated mismatches. The coplanar compensations are chosen so that the plane at which the termination appears purely resistive remains nearly fixed as a function of frequency. Figure 13 shows both the measured variation in the resistance (in percent) of a 100-ohm termination in a 50-ohm system and the measured variation in the effective position relative to the connector reference plane, both as functions of frequency. The effective resistance of the 100-ohm standard is within one percent of 100 ohms from dc through 9 GHz. The effective resistance of the 200-ohm standard is within one percent of 200 ohms from dc through 7 GHz.

Adaptors Between Line Sizes

Transitions

Once precision measuring equipment is available in one line size, accurate measurements in other line sizes equipped with all types of connectors can be achieved with precision adaptors. These adaptors can usually be designed so that their residual reflections are small, and often negligible, relative to the residual reflections of the various connector types.

The principle of the contoured section of transmission line described for the matched termination is easily applied to the design of nearly reflectionless transitions between two line sizes of the same characteristic impedance. An ogee curve is used for the outer conductor with the second half of the curve being a mirror image of the first half. The inner-conductor contour is determined by point-by-point calculation from the outer-conductor contour. The length of the transition section is not fixed, as it was in the termination transition, since the slope at the point of inflection is not defined. However, performance deteriorates as the slope is increased, since eventually the transition approaches an offset step in line size. Transitions with slopes as great as 24° have been fabricated and measured and exhibit essentially no reflections. Such a slope permits a transition from the $\frac{9}{16}$-inch line size to a $\frac{1}{8}$-inch line size in a length of only one inch.

Slot and Gap Effects

A pair of precision adaptors between the $\frac{9}{16}$-inch line size (with precision connectors) and the 7-mm line size, with modified Type N connectors, are shown in cross section in Fig. 14(a). These connectors, modified to introduce minimum reflections, mate with each other and with the military standard connectors in a manner that is mechanically identical with that of the military standard connectors.

The characteristic impedance is constant throughout the modified Type N connectors to satisfy Design Prin-

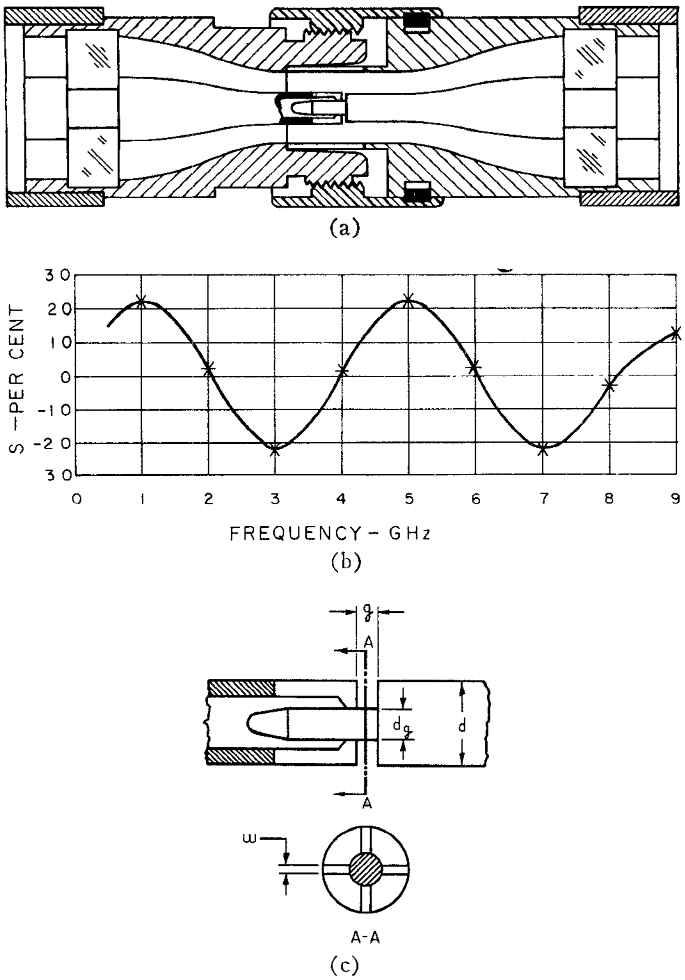

Fig. 14. (a) Cross section of a pair of adaptors to Type N from the 9/16-inch line size; (b) measured effect of axial slots in a reference air line conductor; (c) geometry of an inner-conductor gap.

ciple 1. Therefore, there are no tapers associated with either the inner- or outer-conductor contact fingers once the connectors are mated, and no ridge in the male outer conductor.

The diameter in the region of the inner-conductor slots is increased to provide coplanar compensation (in accordance with Design Principle 2) for the slot effect. The characteristic-impedance error introduced by slots in a 50-ohm air dielectric line (for narrow slots in the inner and outer conductors, respectively) is given by

$$\Delta Z = +12.5N\left(\frac{w}{d}\right)^2, \text{ percent} \tag{9a}$$

$$\Delta Z = +12.5N\left(\frac{W}{D}\right)^2, \text{ percent} \tag{9b}$$

where

ΔZ is the characteristic-impedance change in percent,
N is the number of slots,
w is the inner-conductor slot width in inches,
W is the outer-conductor slot width in inches,
d is the inner-conductor diameter in inches,
D is the outer-conductor diameter in inches.

The conductor diameter changes required for compensation of the slot effects are given directly in terms of the slot dimensions as

$$\Delta d = +104N\frac{(w)^2}{d} \tag{10a}$$

$$\Delta D = -104N\frac{(W)^2}{D} \tag{10b}$$

where Δd is the inner-conductor diameter change in mils, ΔD is the outer-conductor diameter change in mils, and the other parameters are as in relations (9a) and (9b). The relations of (9) and (10) are based on the slot formula given by Montgomery [20].

Figure 14 (b) shows some data measured on a $\frac{9}{16}$-inch reference air line whose inner conductor was slotted. Four axial slots, 0.0354-inch wide and a quarter-wavelength long at one GHz, were introduced to simulate the slot effect of the Type N female inner contact. The measured value of 1.10 percent in characteristic impedance agrees with (9a) within the resolution of the measuring equipment (0.05 percent). Similar measurements using a slotted outer conductor also agree with (9b). The wall thickness of the slotted members, in both cases, was greater than 10 percent of the conductor diameter involved.

In the Type N adaptor pair, there remains the unavoidable discontinuity in the immediate vicinity of the inner-conductor junction. This discontinuity is minimized with the modified connectors by use of a gap of only 2 mils. In addition, axial tolerances contributing to the inner-conductor gap are closely controlled.

The SWR introduced by such a gap depends on both the width of the gap and the width of the slots in the female contact. A formula for this SWR (assuming that the gap forms a radial transmission line that is short compared with a wavelength) is:

$$S = +0.064fg \ln\left(\frac{\pi d - Nw}{\pi d_g - Nw}\right), \text{ percent} \tag{11a}$$

where

S is the SWR in percent,
f is the frequency in GHz,
g is the gap width in mils,
d_g is the inner-conductor diameter in the gap region in inches,
d is the nominal inner-conductor diameter in inches [see Fig. 14(c)].

The remaining variables, N and w, refer to the slots in the female contact, N being the number of slots and w the width of the slots in inches.

The corresponding expression for the SWR introduced by a gap in the outer conductor is:

$$S = +0.064fg \ln\left(\frac{\pi D_g - NW}{\pi D - NW}\right), \text{ percent} \tag{11b}$$

where S, f, g, and N are as before, D_g is the outer-conductor diameter in the gap region in inches, D is the nominal outer-conductor diameter in inches and W is the width of the slots in the outer-conductor contact in inches. As an example, the constants associated with the gap at the Type N inner-conductor junction for the adaptors discussed are $d=0.120$ inch, $d_g=0.065$ inch, $N=4$, and $w=0.016$ inch. With this constant, relation (11a) reduces to

$$S = +\,0.051fg. \tag{12}$$

Thus a gap of 10 mils introduces an SWR of about 3 percent at 6 GHz. Measured data at 6 GHz for gaps from 2 to 40 mils wide agree with relation (12) within 5 percent.

Adaptor Performance

Adaptors with similar connector modifications have also been developed for Type TNC, BNC, and C connectors. Figure 15 shows the SWR performance achieved for various adaptor types.

Although the modified designs of the UG portions are characterized by truly surprising SWR performance for these connector types, the designs of all these adaptors are nevertheless limited by the basic configuration of the UG connectors. Even the modified designs cannot overcome the inherent limitation imposed by dependence of electrical performance on many mechanical tolerances, some of which are subject to wear (violation of Design Principle 3). However, Design Principles 1 and 2 (constant characteristic impedance and coplanar compensation) can be met, and these adaptors are perhaps the best means of interconnecting the various connector types with a minimum of residual reflections.

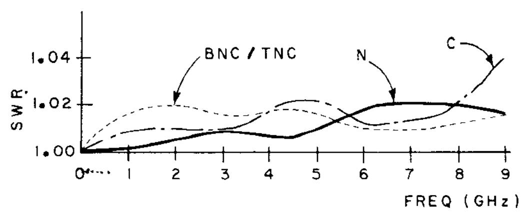

Fig. 15. Measured SWR's of a pair of adaptors to each of Type N, Type BNC/TNC, and Type C connector series from the 9/16-inch line size.

Impedance-Matching Tuner

An impedance-matching tuner is useful and often necessary to tune out the residual reflections of measuring instruments and terminations and to establish initially matched conditions in substitution measurements. To be able to resolve impedance differences that are very small relative to the nominal characteristic impedance, an impedance-matching tuner must be able to introduce very small changes in the degree of match both easily and repeatably. In a precision system, however, the matching range of the tuner need not be especially large.

The impedance-matching tuner shown in Fig. 16 comprises three matching elements spaced at various intervals along a section of constant 50-ohm coaxial line (Design Principle 1). Each tuning element comprises: 1) an adjustable capacitive post protruding through the outer conductor as shown, and 2) an undercut (inductive) region on the inner-conductor coplanar with the adjustable post (Design Principle 2).

When the screw-driven post is all the way out of the region between the conductors, a net inductance is introduced by the tuning element owing to the undercut on the inner conductor. As the post is inserted between the conductors, a capacitance is introduced that offsets the inductance of the undercut. At one position of the adjustable post, its capacitance and the undercut inductance just cancel each other. This neutral position is practically independent of frequency owing to the coplanar nature of the reactances involved (Design Principle 2); therefore, at this position the tuning element is essentially reflectionless, appearing as a simple continuation of the 50-ohm coaxial line. As the post is adjusted beyond neutral, a net capacitance results. Thus, the tuning element introduces both positive and negative reactance relative to the neutral position.

Two such tuning elements, located an eighth wavelength apart at f_0, provide quadrature adjustments on the impedance plane about the 50-ohm characteristic impedance. The angle on the impedance plane between the two adjustments remains within 30° of quadrature over a 2:1 frequency band from 2/3 to $4/3f_0$ and also over a 1.25:1 frequency band centered about $3f_0$, where the tuning elements are spaced by three eighths of a wavelength. This near-quadrature operation results in rapid convergence to match with successive adjustments of the tuning posts. A third tuning element, spaced a different distance from the first, provides a new center frequency f_0, with each of the first two elements taken one at a time. Thus, only two tuning elements are adjusted at any given frequency, but three or more elements can be introduced in a single unit to provide extremely broadband frequency coverage. The unused tuning element(s) at a given frequency is set to neutral where it introduces no significant reflections.

Since the indicated mismatch at the starting point (neutral) is nearly equal to that of the component to be matched and, since the tuning adjustments are nearly orthogonal, convergence to match is rapid.

The tuner shown in Fig. 16 covers the frequency band from 1–9 GHz. The SWR that can be matched out under all conditions of phase increases linearly from 1.012 at 1 GHz to 1.10 at 9 GHz. The residual SWR introduced by the tuner when all posts are set to neutral also increases linearly at about half this rate. Because the distance between the extreme tuning elements need be only about 0.09λ at the lowest frequency of interest, the size of this tuner, complete with connectors and calibrated scales, is only 6 inches×4 inches×1 inch. Figure 17 shows the measured tuning range over the octave

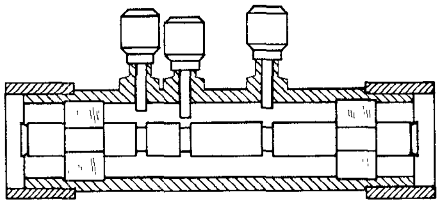

Fig. 16. Schematic cross section of a three-element impedance-matching tuner.

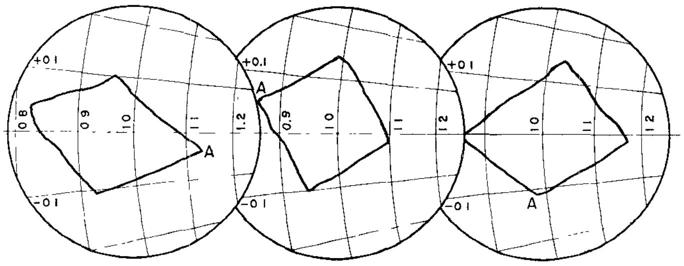

Fig. 17. Measured matching range of a two-element tuning section centered about $f_0=5.4$ GHz.

from 3.5 to 7 GHz for the two tuning elements spaced 0.7 cm apart (one-eighth wavelength at 5.4 GHz).

Conclusions

Precision coaxial connectors and the instruments, standards, and components built around them comprise a branch of microwaves that is relatively new, but one that is growing rapidly because of increased industry demand for systems with high performances over dc-to-cutoff-frequency bandwidths. The advances in performance presented in this paper substantiate a design philosophy for precision devices based upon the elimination of impedance discontinuities on an incremental basis. These design principles, if followed rigorously, allow the design engineer to converge on the optimum in accurate broadband standards and near-reflectionless broadband components.

It is anticipated that the advances noted here will, in turn, generate advances in other impedance standards and components and in more sophisticated microwave measuring instruments.

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