STRIP TYPE COMPONENTS FOR 2000 MEGACYCLE RECEIVER HEAD-END

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Summary

Recent experimental work has evolved some components using air-spaced strip type transmission line that have been used successfully in connection with a variable attenuator, cavity and crystal mixer. Bandwidth, VSWR, and NF measurements are comparable with a commercial receiver head-end presently in use. The asymmetric air-spaced strip-above-ground transmission line used results in a simple configuration for coupling the line to the cavity. It also facilitates the application of a variable attenuator using a ferrite slab of high attenuation per unit length and having good VSWR properties.

I. Introduction

Experimental investigations were undertaken to study the possibility of adapting a presently used 2000 mcs receiver head-end to microstrip type components. The factors which initiated this investigation were to reduce the physical size as well as the production cost of the head-end without sacrificing the electrical properties.

Several papers,[1,2,3,4,5] have presented a descriptive and experimental viewpoint of microstrip components as well as a first-order theoretical analysis of the transmission properties of these lines.

The type of microstrip transmission line described in this paper is of the simplest kind, representing an asymmetrical air-spaced strip-above-ground transmission line. The reason for choosing this particular type of line was the fact that provision had to be made for a variable attenuator of high attenuation ahead of the input cavity. A transmission strip line of the printed type, that is, with dielectric support for the strip conductor, in connection with a slab attenuator, produces rather low attenuation values per unit length[2] when the attenuator is placed on top of the strip line. The air-spaced strip line, on the other hand, allows the insertion of suitable attenuating material between the conductor and ground plate yielding high attenuation per unit length.

The receiver head-end included the following components: input connector, a short length of strip line, variable attenuator, tunable cavity, a short length of strip line, variable local oscillator signal injection, and crystal mixer. The additional element which had to be incorporated in comparison to the standard head-end was the variable attenuator which determined the kind of strip line used as mentioned previously.

II. Discussion

The input connector had to be separated from the cavity in order to have a long enough strip-above-ground transmission line in which sufficient attenuation could be inserted. A suitable Ferrite material with high attenuation per unit length was found to work satisfactorily.

A strip line with an impedance of essentially 50 ohms, made of 1/16" copper and 5/16" wide, with 1/16" spacing from the ground plate was used as a standard element and showed negligible insertion loss for short line lengths.

The transition from a type "N" flange connector to the strip-line can be made, either with or without a quarter-wave short-circuited stub depending on the required mechanical rigidity of the line. A short open stub resulted in a very good match over the frequency range and is less frequency sensitive than the transition with the short-circuited quarter-wave stub.

The application of a resonant strip line in place of the cavity was studied and found to have too much radiation loss unless the shielding was elaborate in which case a cavity of simple design gave a less complex configuration and showed a smaller insertion loss.

Any aperture in a 1/16" thick plate mounted at right angles to the strip line and normal to the ground plane with an opening so that the remaining air gap between the strip conductor and the aperture is at least half of the strip width, has no appreciable effect on the propagation characteristics.

The injection of the local oscillator signal was accomplished with a rotatable loop or probe at such a position that the RF signal is not coupled back to the local oscillator.

The mixer crystal can be mounted in two different ways depending on the layout of the head-end components with respect to the IF amplifier: a) with the RF bypass at the crystal head and the IF signal fed off from the same point, or b) by using a coupling capacitor at the crystal pin with the head-end dc grounded and the IF taken off from the pin by means of a quarter-wave RF choke.

It was noticed that the short quarter-wave stub for the crystal had to be longer than a quarter wavelength for matching purposes. Since the space requirements didn't allow for such a long stub an open stub of short length was used which eliminated at the same time the possibility of taking off the IF signal from the crystal pin. The variation of the band-width and VSWR over tuning range stayed within the specifications.

Comparative NF measurements between the two head-end types showed no difference as long as the RF bypass capacity of the two crystal holders and the bandwidth of the cavities had approximately the same values.

III. Conclusions

It is feasible to replace the present head-end with one using construction incorporating short sections of the asymmetric strip-above-ground transmission line. This head-end is simpler mechanically and less expensive than the original head-end. A deterioration of the electrical properties, in particular the NF, could not be detected.

IV. Electrical Properties of Strip-Type Components

The various microstrip components and the corresponding experimental data will be given in the order in which they were investigated.

1. Air-spaced strip-above-ground transmission line and transitions to type "N" connectors (UG-58/U)

The impedance properties of the asymmetrical air-spaced strip-above-ground transmission line of finite thickness have been given[6] as a function of the ratio of strip width b to strip spacing h from the ground-plate and with the ratio of strip thickness d to strip width as parameter. The approximate expression of the strip line impedance (d small) is:

$$Z_0 = Z_0' \frac{1}{2 + \frac{b}{h} + \frac{1}{\pi}\frac{d}{h}\left[1 + \ln\left(1+\frac{2h}{d}\right)\right]} \qquad (1)$$

$$\text{with } Z_0' = \sqrt{\frac{\mu_0}{\varepsilon_0}} \qquad (2)$$

A strip-line impedance of essentially 50 ohms was of importance in order to reduce the mismatch when the line is used in connection with type "N" connectors.

The impedance is essentially controlled by the ratio (b/h) and to a smaller extent by the parameter ratio (d/b). Fig. 1 shows the plot of the impedance function for (d/b) = 0.20 and for a strip of zero thickness d = 0.

For the strip of zero thickness above an infinite ground plane the approximate expression (b/h large) derived by F. Assadourian[3] was applied:

$$Z_0 = Z_0' \frac{1}{\frac{b}{h} + \frac{2}{\pi}\left[\ln\left(1+\frac{\pi b}{2h}\right) + 1\right]} \qquad (3)$$

This is represented by the straight line in Fig. 1. The difference in impedance between the two curves is in the order of 10% for that particular ratio d/b = 0.2.

One of the first questions which arises when using an air-spaced strip-above-ground line is: How important is the preservation of proper spacing, or what variation in spacing can be tolerated for a specified mismatch? This question is of secondary importance for a printed line with solid dielectric. From Eq. (1) with b/h = 5 and d/b = 0.2, which represents a characteristic impedance of essentially 50 ohms, it follows that for a spacing reduction of 10% the impedance increases by 7%.

The dimensions of the copper strip line were chosen as follows:

b = 5/16"   h = d = 1/16"

This resulted in a rather narrow strip line which had, however, sufficient rigidity for a short line length. The properties of this line were measured in connection with two type "N" flange-connectors (UG-58/U) mounted on a 1/8" thick copper ground plane 3-1/2" wide.

The impedance properties of two lines were measured, a 4" long line with quarter-wave short-circuited stubs, adjusted for minimum VSWR at 1800 mcs and a line 8" long with open stubs of 3/16" length as shown in Fig. 2. Tables I and II summarize these results.

Table I

Line 4" long with short-circuited quarter-wave stubs at each side. Stub lengths 42mm adjusted for minimum VSWR at 1800mcs.

Frequency mcs	Load VSWR	Line and Load VSWR
1700	1.08	1.12
1800	1.08	1.05
1900	1.08	1.16
2000	1.08	1.30

Table II

Line 8" long with open stubs of 3/16" length.

Frequency mcs	Load VSWR	Line and Load VSWR
1700	1.08	1.08
1800	1.08	1.11
1900	1.08	1.16
2000	1.08	1.20

No steps were undertaken to compensate the increased VSWR at the high frequency side since the deterioration is only small.

The stub length of 42 mm represents practically a quarter-wave length at 1800 mcs, that is, the frequency at which the stub length was adjusted for minimum VSWR. This would indicate a very small susceptive component of the transition.

The insertion loss of the 8" long strip line including the transitions didn't exceed 0.2 db over the frequency band.

The short open stub can be used advantageously for matching purposes in connection with other components of the head-end as will be shown in paragraph 3.

2. Asymmetrical air-spaced strip-above-ground resonant line

In order to determine the possibility of replacing the cavity by a resonant strip-line configuration, the following measurements were made.

A strip, one half-wavelength long, shorted at both ends, was very loosely coupled to two small rotatable loops mounted on UG-625/U connectors, threaded in the ground plate and placed near the short-circuited ends.

The half-power bandwidth was then measured for various strip widths and spacings from the ground plate. The application of strip-line components as resonant elements is not new. High-Q strip-line elements have been described elsewhere.[7] A symmetrical construction was used and it was pointed out that the preservation of symmetry of the structure was essential in order to avoid the introduction of spurious modes. This would automatically complicate the tuning mechanism for a bandpass filter of large tuning range. The measurements were therefore limited in this case to the asymmetrical resonant strip-line.

The strip was made of 1/16" copper, soldered to a ground plate 7" x 12" and the length inside was a half wavelength at 2000 mcs. The strip spacing was 1/16" from the ground plate.

Fig. 3 represents the change in unloaded Q which was computed from the bandwidth measurement and the shift of the resonant frequency with spacing. The radiation along the open sides of the strip was quite pronounced and increased for greater strip spacing.

The term unloaded Q may be misleading here since the Q-values are so low. The loading, however, is not due to the generator and load impedance but determined by the strong radiation of the resonant structure. The unloaded Q increases with smaller spacing and the resonant frequency moves towards lower values.

Different configurations were examined to reduce the radiation. The most promising simple layout was the half-wave resonant strip described above with two shields 2" high and 5" long placed parallel to the resonant strip and symmetrically spaced.

Fig. 4 shows the improvement in Qu due to the shields and the variation of the resonant frequency.

The spacing was not critical and approximately a quarter-wavelength wide. The Qu values increased from 96, without the shields, to 580 for optimum shield spacing s. The resonant frequency dropped from 1932 mcs, without the shields, to 1927 mcs for optimum s. The radiation was greatly reduced; a shorting bar moved along the top of the shields had no effect. Cutting down the height of the shield to 1", however, increased the radiation. The same maximum Qu value could be achieved by enclosing the resonant strip in a box 4" x 1-1/2" x 1/2" which would indicate that the loss resistance of the resonant strip was a contributing factor.

This resonant strip without the coupling loops, but with the shields, was then coupled to two strip lines 2-1/2" long with quarter-wave stub supports at the transitions to type "N" flange connectors. A 1/2" No. 32-thread brass tuning screw was threaded in the ground plate at the center of the resonant strip. The input-and-output strip lines were coupled to the resonant strip by soldering, near the short-circuited ends and parallel to the resonant strip line, a 5/16" long part of the strip line. Fig. 5 illustrates this coupling configuration.

The two shields were flared at the ends to make room for the strip lines. Table III summarizes the measured values.

Table III

Tunable resonant line with shields 5" x 2". Stub lengths 30 mm, adjusted for min VSWR at 1900 mcs.

Frequency	1800	1900	2000
VSWR	1.11	1.06	1.16
VSWR of Load Alone	1.14	1.10	1.10
Insertion Loss, db	1.2	1.4	1.2
Bandwidth, mcs	23.6	23.2	23.5
Loaded Q, Q_L	76.3	82.0	85.0
Unloaded Q, Qu	640	595	710

The unloaded Q, Q_u was calculated from the expression of the insertion loss. For matched conditions:

$$G_T' = -10 \log(1 - 2\, Q_L'/Qu), \text{db} \qquad (4)$$

which can be derived from the general expressions:

$$1/Q_L = 1/Q_1 + 1/Q_2 + 1/Q_u \qquad (5)$$

and

$$G_T = 20 \log 1/2[\sqrt{Q_1/Q_2} + \sqrt{Q_2/Q_1} + \sqrt{Q_1 Q_2/Qu}], \text{db} \qquad (6)$$

by differentiating Eq. (6) with respect to Q_1 for a given Q_2 or vice versa. Q_1 and Q_2 are the external Q's of the input respectively output circuit and are functions of input and output quantities only.

The insertion loss still represented a rather high value. The figure in mind was less than 1 db. Higher unloaded Q-values could certainly be obtained by changing the resonant-strip configuration. It was, however, unattractive to carry out these investigations further since some kind of shielding would have been necessary in any case in order to suppress the radiation loss. Therefore, it was decided to replace the resonant half-wavelength strip with a cavity of simple design and physically small size.

Various kinds of coupling possibilities from the strip line to the cavity became evident. The simplest form which uses the air-spaced strip-above-ground transmission line itself, or any shaped portions of it, as a coupling loop or probe inside the cavity was investigated in more detail. This resulted in the following study.

3. <u>Air-spaced strip-above-ground transmission line and aperture in plane at right angle to the direction of propagation</u>

The change in strip-line impedance due to a plane of finite thickness with an aperture of specified size placed over the strip line at right angle to the direction of propagation and normal to the ground plate was measured in order to determine the quantitative effects.

The strip line used was 8" long with open stub transitions as described in paragraph 1. The square plate (3" x 3") was placed at a half-wavelength (2000 mcs) from the pin of the type "N" flange connector. The effect of three different kinds of apertures cut in a 1/16" thick plate were measured: a) 3/4" half round hole, b) 1/2" half round hole, both centers of the half circles were spaced 1/16" from the ground plate, and c) rectangular opening 7/16" x 3/16". This last aperture which evidently would produce the greatest influence was also cut in a 3/16" thick plate.

The susceptance component is capacitative and increases with frequency and smaller aperture sizes. The measured values do not represent the absolute susceptance of the aperture alone but include the transition from strip line to type "N" connector. One can, however, get a quantitative picture of the aperture effect. For example, increasing the plate thickness from 1/16" to 3/16" for the rectangular aperture has a smaller effect than replacing the 1/2" half round aperture by the rectangular aperture 5/16" x 3/16".

The measured values of the aperture susceptance are characteristic only for this half-wavelength spacing of the aperture plate from the type "N" connector. Since the short open stub at the transition represents also a capacitive susceptance its value can be applied successfully to compensate the capacitive susceptance component of the aperture which would become inductive for quarter-wavelength spacing from the transition point. Two parameters are therefore available: a) the distance of the aperture from the transition point and b) the length of the short open stub. A shortened short-circuited quarter-wavelength stub can be used on the other hand for a half-wavelength spacing of the aperture plate from the transition point.

A second series of measurements were made to get a more sensitive indication of the aperture effect. The strip line was shorted and the phase displacement of the voltage minimum of the VSWR observed when the aperture plate was inserted at a voltage maximum, that is, a quarter-wavelength from the short-circuited point of the strip line. The distance from the plate to the pin of the type "N" connector was a half-wavelength.

Table IV summarizes the measurement results.

Table IV

Phase displacement of the voltage minimum of the VSWR caused by the aperture plate inserted at a voltage maximum. Frequency 2000 mcs.

Phase displacement of the voltage minimum of the VSWR		$\Delta l/\lambda$
3/4" half round aperture	in 1/16" thick plate	-0.0133
1/2" " " "	in 1/16" thick plate	-0.020
7/16" x 3/16" rectangular aperture	in 1/16" thick plate	-0.033
7/16" x 3/16" rectangular aperture, in 1/8" plate		-0.033

The aperture effect is not pronounced for large apertures and for small apertures not a function of the plate thickness. No appreciable effect on the propagation characteristic can be detected as long as the remaining air gap between the aperture and the strip line is at least half the width of the air-spaced strip-above-ground line.

4. Circular cavity coupled to air-spaced strip-above-ground transmission lines

The inside dimensions of the circular cavity were as follows: 1-7/8"ID, 1-1/8" high, wall thickness 1/16". Two half round 1/2" diameter holes were used for the input and output coupling. The strip line inserted through the aperture and shorted to the ground plate so that the part inside the cavity formed the coupling loop. The spacing of 1/16" from the ground plate was retained and the coupling determined by adjusting the loop length inside the cavity. (See Fig. 7)

The tuning was accomplished by a 1/2" diameter No.32-thread brass screw which loaded the cavity capacitively. The tuning range from 1700 to 2000 mcs was covered with 2-1/2 turns.

This kind of coupling configuration can be applied with advantage wherever a tight coupling, that is a relative bandwidth of more than 1%, is required. The placement of the cavity and the input and output strip line on the same side of the ground plate results in a shielding effect due to the cavity between the two strip lines.

Since the bandwidth of the cavity is affected to some extent by the length of the quarter-wave short-circuited stubs at the transitions, two series of measurements were taken: a) transition with short-circuited stubs, two different stub lengths and b) transition with open stubs.

No particular length was chosen for the distance between the cavity outside wall and the transition point (2-1/2") in order to compensate the aperture effect for this coupling configuration.

Table V

VSWR, bandwidth, loaded Q, insertion loss, and unloaded Q as function of frequency for circular cavity coupled to air-spaced strip-above-ground line.

a) Short-circuited stubs, 40 mm long

Frequency, mcs	1800	1900	2000
VSWR	1.14	1.20	1.14
VSWR of load alone	1.17	1.13	1.12
Bandwidth, mcs	24.3	25.6	28.6
Loaded Q	74.2	74.2	70.0
Insertion loss, db	0.7	0.8	0.7
Unloaded Q	1000	885	940

b) Short-circuited stubs, 30 mm long

Frequency, mcs	1800	1900	2000
VSWR	1.06	1.20	1.04
Bandwidth, mcs	25.0	23.2	24.4
Loaded Q	72	82	82
Insertion loss, db	0.7	0.9	0.7

c) Open stubs

Frequency, mcs	1800	1900	2000
VSWR	1.19	1.25	1.03
Bandwidth, mcs	23.6	27.2	32.2
Loaded Q	76.4	70.0	62.0
Insertion loss	0.8	0.8	0.9

A shift of the resonant frequency from 1900 to 1905 mcs was observed with the reduction of the stub lengths from 40 to 30 mm. The increase in bandwidth at 2000 mcs for the case with open circuited stubs can be explained by the fact that the resonant frequency shifts to a lower value when the short-circuited stubs are replaced by open circuited stubs, that is, in order to tune to the same frequency again the tuning screw has to be turned out which changes the field distribution inside the cavity and therefore increases the coupling to the strip lines. The position of the tuning screw and its effect on the coupling becomes more important for cavities which are highly capacitively loaded and of small height as in this case.

This same effect was observed to a more aggravated extent for a square cavity, 3" x 3" and only 1/2" high, capacitively loaded with the same tuning screw. The coupling configuration, however, was different. The cavity and the coupling strip lines were placed on opposite sides of the ground plate and the coupling achieved by means of a 1/2" diameter coupling hole in the ground plate, the end of the strip screwed to a coupling post which was inserted through this hole into the cavity and fixed inside to its top.

The unloaded Q was computed from Eq. (4). The insertion loss stayed below 1 db over the frequency band.

A different cavity of square cross-section and substantially greater height was used for the final measurements in order to get less variation of the loaded Q over the frequency range. This resulted simultaneously in a further reduced insertion loss.

5. <u>Square cavity coupled to air-spaced strip-above-ground lines with crystal mixer and local oscillator injection</u>

This series of final measurements was made on the head-end complete, except for the variable attenuator.

a) <u>Square Cavity</u>

The cavity had a cross-section of 2" x 2" and was 2-1/2" high. The same tuning screw and coupling configuration were applied to this cavity as described for the circular cavity. The tuning range from 1700 to 2000 mcs was covered with 8 turns of the tuning screw, indicating a much smaller capacitive loading than for the circular cavity. The bandwidth measurements were carried out with transitions of short-circuited stubs, both 28 mm long. The length of the strip line was the same as for the circular cavity.

Table VI

VSWR and bandwidth as function of frequency for the square cavity coupled to air-spaced strip-above-ground line. Short-circuited stubs, 28 mm long.

Frequency, mcs	1700	2000
VSWR	1.12	1.26
Bandwidth, mcs	22.2	23.9
Loaded Q	76.5	83.6

The insertion loss at 1700 mcs was 0.56 db. The variation in loaded Q was now less than 10% and proved to be satisfactory.

b) <u>Local Oscillator Injection</u>

The simplest form of L.O. injection was accomplished by a rotatable loop mounted on a UG-625/U connector, threaded in the ground-plate. The available L.O. power was sufficient to allow decoupling of the loop from the strip line which assured negligible loss of the RF signal. Spacing the loop approximately a quarter-wavelength from the cavity and the crystal mixer resulted in the best decoupling since the line towards the cavity represents then a high impedance at the L.O. frequency. The strip line length for this L.O. injection configuration required, however, too much space. Different ways of L.O. injection can be thought of which would represent, for example, a more uniform impedance for the L.O. signal over the tuning range. A strip line type coupling configuration with a variable capacitive probe is also possible where the space requirements are not to strict.

c) <u>Crystal Mixer</u>

The crystal mixer was of conventional design, except for the fact that the IF lead had to be fed through the ground plate since the crystal had to be easily accessible from the front side, the same side from which the tuning of the cavity was achieved. The IF strip was mounted on the ground plate in such a way that the IF input terminal fell on the same side as the strip line.

The half round 1/2" diameter coupling holes in the cavity were cut in adjacent walls in order to reduce the head-end size and still have a sufficiently long input strip line for the insertion of the variable attenuator.

Fig. 8 shows the layout of the various components.

The output line to the crystal mixer was fairly short but long enough to allow room for the L.O. coupling loop. The transition to the crystal made use of an open circuited stub.

A dielectric support for better mechanical rigidity would be necessary at the open side of the stub. The complication of feeding the IF lead through the ground plate can be avoided by using a coupling capacitor at the crystal pin. This would call for a short-circuited quarter-wavelength strip-line-stub along whose top surface the IF lead could be taken off. Since the space requirements were of importance, it was not possible to pursue this solution.

Table VII

VSWR, bandwidth and loaded Q of receiver head-end as function of frequency. Crystal current 0.4 mA.

Frequency, mcs	VSWR	Bandwidth, mcs	Loaded Q
1740	1.18	20.0	87.0
1800	1.20	21.1	85.3
1900	1.26	22.0	86.5
2000	1.49	23.8	84.0

The bandwidth, with the crystal as the loading element at the output side, was measured for constant frequency of the IF signal, varying the frequency of the input and L.O. signal, and keeping the crystal current constant.

Comparative NF measurements between a conventional, machined receiver head-end and the one above described didn't reveal a difference for similar operating conditions. Shielding of the strip line type-head was necessary when making these measurements.

6. Variable Attenuator

The three important factors which control the attenuation can be classified as follows: a) type of material, b) dimension of the attenuating element, and c) condition of the surfaces.

Since the attenuator had to be inserted between the strip line and the ground plate in order to obtain high enough attenuation values, the height of the element was automatically fixed for a strip line of specified impedance and width. The only remaining parameter with regard to the dimensions of the element was its length. The width didn't contribute a great deal to the attenuation as long as the element was somewhat wider than the strip line width. A relatively short attenuating element was requested in order to cut down the length of the input strip line. The factor contributing most to the attentuation was consequently the type of material of the element.

The attenuating properties of several materials were investigated until a suitable element in the ferrite group was found.[8,9,10] Two types of ferrites, a Ni-Zn and Mn composition were measured, the latter having a 10 to 20 db higher attenuation for the same sample length. The Curie point temperature was well above 100°C assuring good thermal stability of the attenuator. The effect on the attenuation of an air gap between the attenuator and the strip line was examined.

The deleterious effect of the air gap on the attenuation as well as the fact that any gap variation would influence the attenuation, made its elimination advisable. A conductive silver film was therefore applied to the sample surfaces in order to achieve good electrical contact with the strip line and the ground plate.

The fired-on silver film, which was in the order of .001", increases the mismatch. By removing the silver film at the ends of the sample or by shaping the film, the VSWR can be greatly decreased. This method of matching is preferable to tapering the element because it would involve an additional machining step.

Table VIII

Effect of the silver film on the VSWR and the attenuation for a Ni-Zn sample 45 x 15 mm at 1800 mcs.

Sample surface	VSWR	Attenuation, db
No silver film	2.0	32
Both sides fired-on silver	3.0	39
Bare ends both sides:		
4 mm wide	1.36	36
7 mm wide	1.50	36
10 mm wide	1.60	36

For both ferrite compositions, the attenuation increases linearly with sample length above a minimum length of 1-1/2 to 2 times the strip width.

No gain in attenuation results by widening the sample of a fixed length by more than 1-1/2 times the strip width.

Table IX

Attenuation as function of sample width. Mn ferrite sample, 14 mm long, at 1700 mcs.

Sample width, mm	Attenuation, db
8 = strip width	34
11	42
12	42
15	42

The attenuation is a function of frequency and increases generally with frequency over the range 1700 to 2000 mcs. The impedance of both ferrite compositions is almost purely resistive. The reactive part being slightly inductive or capacitive, depending on the sample length. For the Ni-Zn samples, the resistive part is almost twice as high as for the Mn samples, resulting in a better match to the strip line impedance.

A gradual increase of the attenuation can be obtained by inserting the ferrite element parallel to the strip line or rotation.

The difference in attenuation for that particular Mn sample between 1700 and 2000 mcs was approximately 10%. A minimum attenuation has to be specified for this particular attenuator shape since the attenuation is discontinuous up to the point where the element comes in contact with the strip line.

In order to expand the insertion distance, the length of the attenuator should be chosen so that the maximum specified attenuation occurs when the element is fully inserted under the strip line. For a composition which has high attenuation per unit length this would call for a rather short sample. The taper of the silver film, which was applied to the top surface only for that particular sample (the bottom side had 3/16" wide bare ends), would consequently have a steep rise and a higher mismatch would have to be tolerated.

Finally, it has to be mentioned that the firing schedule in calcination and sintering have to be specified during the manufacturing process of the ferrite in order to obtain reproducible results.

V. Acknowledgment

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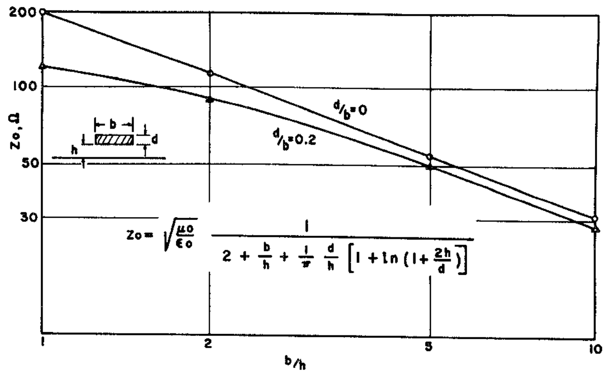

Fig. 1 - Characteristic impedance Z_o of asymmetrical strip-above-ground transmission line. Parameter d/b.

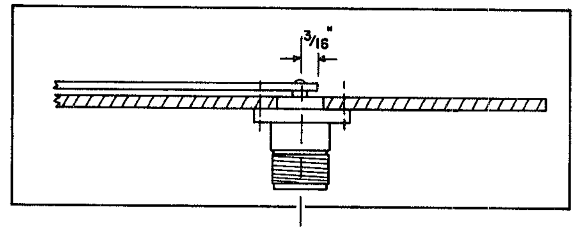

Fig. 2 - Transition from type "N" flange-connector (UG-58/U) to air-spaced strip-above-ground transmission line.

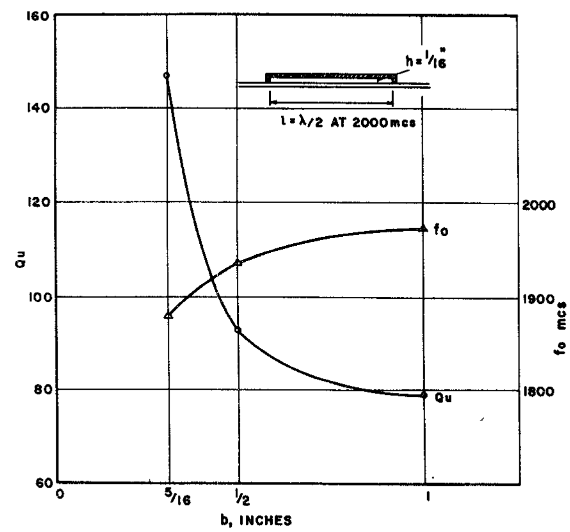

Fig. 3 - Unloaded Q and resonant frequency as function of strip width b.

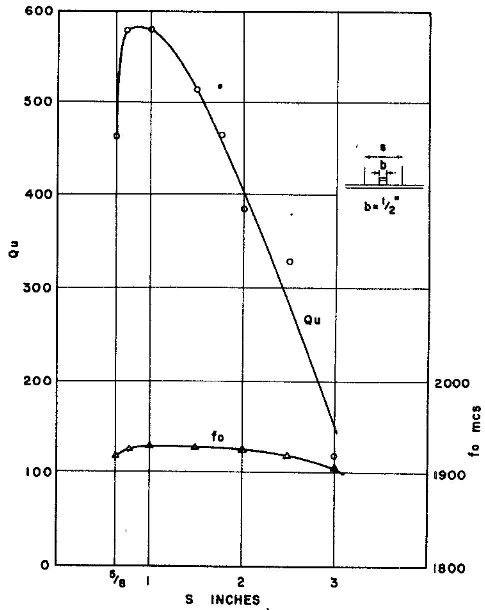

Fig. 4 - Unloaded Q and resonant frequency as function of shield spacing S.

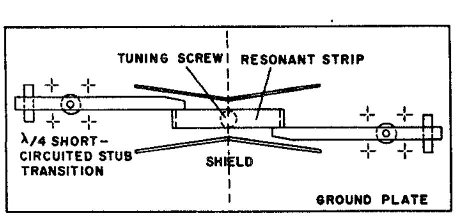

Fig. 5 - Coupling configuration of strip line to half wave length resonant strip.

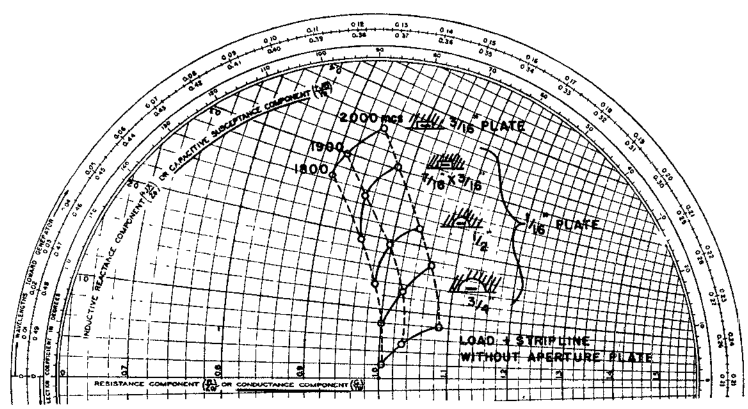

Fig. 6 - Susceptance of aperture for various shapes and sizes as function of frequency.

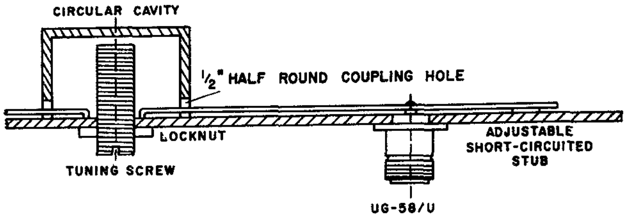

Fig. 7 - Circular cavity coupled to air-spaced strip-above-ground line.

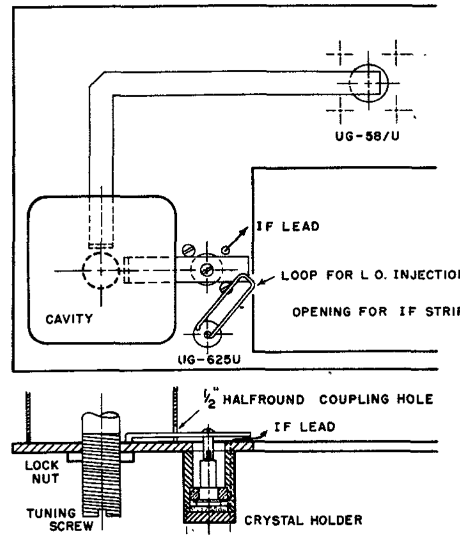

Fig. 8 - Receiver head-end with air-spaced strip-above-ground coupling lines.

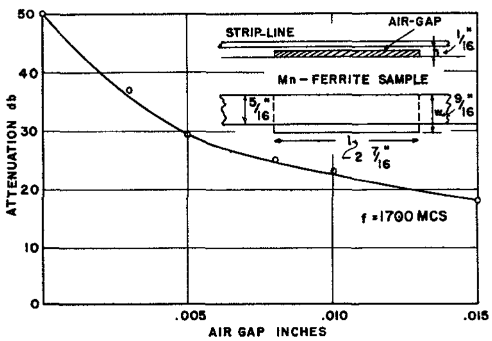

Fig. 9 - Attenuation as function of air-gap between sample and strip line.

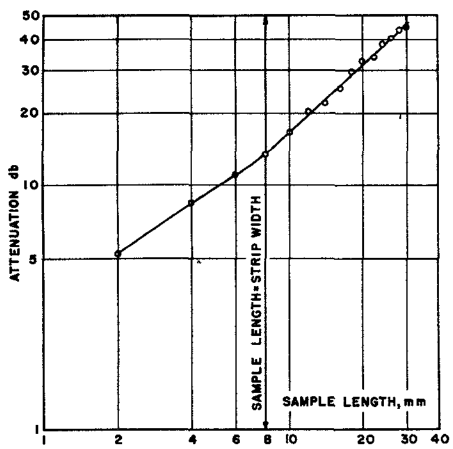

Fig. 10 - Attenuation as function of sample length for a Mn ferrite at 1700 mcs. Sample width = strip width.

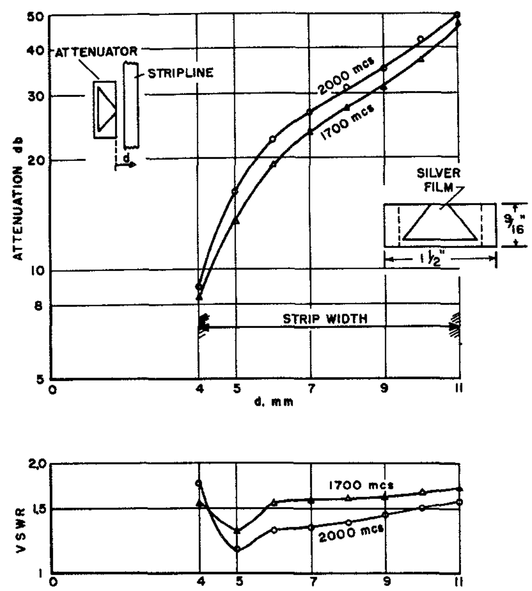

Fig. 11 - Attenuation and VSWR as function of attenuator position. Parameter : frequency.