A Network for Combining Radio Systems at 4, 6 and 11 kmc

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Development of the broadband horn reflector antenna has permitted the simultaneous radiation and reception of radio signals on different frequencies in the three common-carrier bands in which the Bell System has developed radio relay systems. A necessary adjunct to the antenna is a network to combine or separate the common carrier bands and also to combine or separate the two polarizations of any one band. The particular form of the network that is described was designed to meet strict system requirements on impedance match, insertion loss and cross-coupling between ports.

Microwave radio relay systems have been developed for Bell Telephone System toll transmission in the common carrier frequency bands of 3700 to 4200 mc, 5925 to 6425 mc and 10,700 to 11,700 mc. There will be many routes where two, or all three, of these radio systems will be in use simultaneously. On such routes, a common antenna[1] at the top of the station tower will transmit or receive all three frequency bands. The broadband antennas will operate with both polarizations of the radio waves, and the feed line will be a 3-inch circular copper waveguide (WC 281). The 3-inch waveguide is capable of supporting three modes in the 4-kmc frequency range and 22 modes in the 11-kmc frequency range. With the possibility of so many propagating modes, small discontinuities in the internal dimensions of the waveguide may cause undesirable mode conversion with associated transmission loss or reflection of signal energy. In particular, any bend made in the circular waveguide must have very small curvature to minimize coupling between the two cross-polarized dominant modes at 11,000 mc.

The cross-coupling requirements mean that the circular waveguide must run nearly straight down the side of the tower to a network located at the base of the tower. After the signals are in dominant-mode waveguide, bends and twists can be utilized to bring the signal into the repeater stations. The systems combining network is required to distribute the six signals (two polarizations at each of three frequency bands, 4, 6 and 11 kmc) to the proper ports with small insertion loss (approximately $\frac{1}{2}$ db attainable), with small reflections (30 db return loss) and with small delay distortion (echoes 60 db below the signal).

The basic configuration of this systems combining network was suggested by Miller,[2] and an experimental network was reported on by Dawson.[3] A series of directional couplers is used to successively distribute the different polarizations and frequency bands. The circular waveguide from the antenna is tapered down to a dominant-mode square waveguide in the 3700- to 4200-mc band. The directional couplers are constructed by running rectangular dominant-mode waveguide parallel to the two-polarization square waveguide, with the narrow face of the rectangular waveguide adjacent to the two-polarization waveguide (see Fig. 1). Longitudinal slots cut in the common wall will couple to only

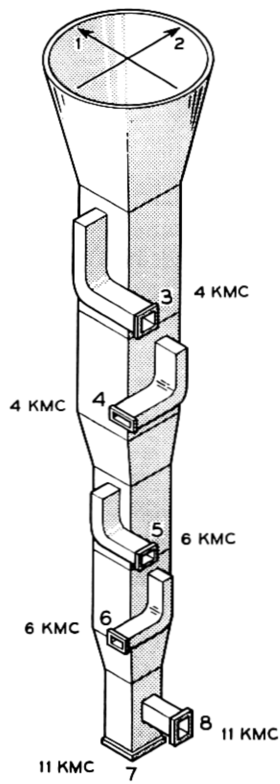


Fig. 1 — Designation of dominant-mode terminal pairs for systems combining network.

one polarization of the 4-kmc signal in the two-polarization waveguide, and a sufficient number of slots will be used to extract all the energy in this polarization. (The frequency-selection means will be discussed later.) A similar directional coupler rotated by 90 degrees is connected below the first coupler and couples to the other polarization of 4 kmc. All the 4-kmc energy in both polarizations has now been extracted, so the two-polarization waveguide is tapered down to square dominant-mode size at 6 kmc. Two more directional couplers with longitudinal slots are connected to extract the two polarizations at 6 kmc. The two-polarization waveguide is again tapered down in size, and a network to combine both polarizations at 11 kmc is connected.

All the directional couplers are reciprocal networks, so the structure can be used to combine the frequencies and polarizations into a single waveguide or to separate the signals from a single waveguide into their respective ports — or various combinations of combining and separating (transmitting and receiving) can be achieved.

The two directional couplers extracting the 4 kmc energy must, of course, not produce appreciable mode conversion of 4-, 6- or 11-kmc energy. Any size waveguide chosen as the two-polarization waveguide for the 4 kmc couplers will propagate several higher-order modes with cutoffs below 11,700 mc. It has been found that the most troublesome regions for higher-order mode couplings are in the vicinity of the mode cutoffs. For this reason, it is desirable to have all the mode cutoffs as far away as possible from the frequency bands 3700 to 4200 mc, 5925 to 6425 mc and 10,700 to 11,700 mc. This selection of mode-cutoff patterns leads to a very restricted choice of waveguide sizes, and results in dominant mode cutoff for the two-polarization waveguide (3300 mc) being close to the lower edge (3700 mc) of the 4-kmc band.

Fig. 2 graphically illustrates the relationship between the mode cutoff pattern and the common-carrier frequency bands. It is clear that the waveguide size or mode cutoff pattern cannot be shifted very much in either direction. The use of longitudinal slots (to separate polarizations) and waveguide operating close to cutoff (to avoid couplings to higher-

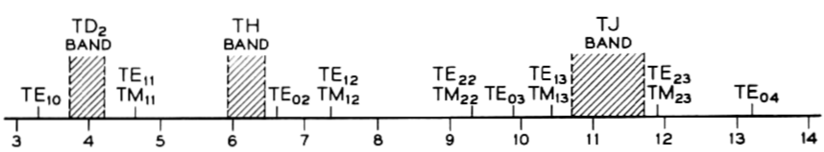


Fig. 2 — Distribution of mode cutoffs with relation to common carrier bands for square waveguide 1.790 inches on a side.

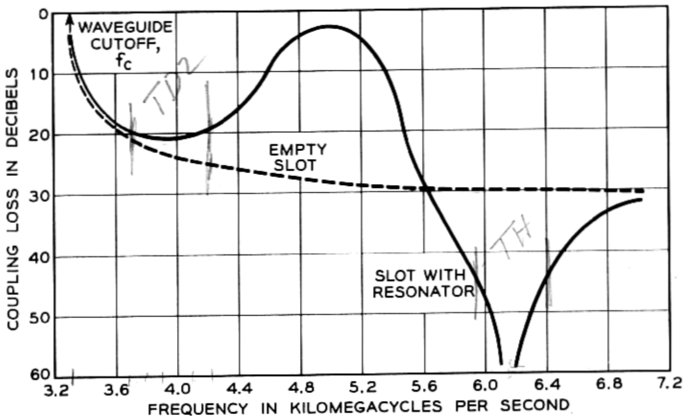


Fig. 3 — Transfer of energy between guides for single slot.

order modes) results in the coupling-versus-frequency characteristic shown in Fig. 3.

Using procedures described by Bethe[4] and Surdin,[5] we have equations for the amplitude of the wave coupled through a small slot in the side wall of a waveguide and the power carried by waves of unit amplitude in each waveguide:

$$A_1 = j \frac{2\pi}{\lambda_0 S_1} M H_{1z} H_{2z} ,$$

$$S_1 = \frac{a_1 b_1 \lambda_{g1}}{2\lambda_0} ,$$

$$S_2 = \frac{a_2 b_2 \lambda_{g2}}{2\lambda_0} ,$$

where

A_1 is the amplitude of the coupled wave,
$S_1 A_1^{\,2}$ is the power carried by a wave in waveguide 1,
S_2 is the power carried by a unit-amplitude wave in waveguide 2,
λ_0 is free-space wavelength,
λ_g is guide wavelength,
H_z is the magnetic field at the coupling slot for a wave of "unit" amplitude,
a and b are waveguide width and height and
M is the magnetic polarizability of the slot in the z direction.

The coefficient of coupling, c, is the square root of the ratio of the power excited in waveguide 1 to the exciting power in waveguide 2:

$$c = \frac{A_1\sqrt{S_1}}{\sqrt{S_2}} = M\,\frac{\pi}{2}\sqrt{\frac{\lambda_{g1}\lambda_{g2}}{a_1^3 b_1 a_2^3 b_2}}. \tag{1}$$

This expression shows that energy coupled between waveguides by a single slot decreases as the frequency is increased across the 4-kmc band.

For an array of coupling apertures, Miller[2] has shown that the amplitude of the wave coupled between two waveguides is given by

$$E_2 = \frac{j}{\sqrt{\left(\frac{B_1 - B_2}{2c}\right)^2 + 1}} \sin\left[nc\sqrt{\left(\frac{B_1 - B_2}{2c}\right)^2 + 1}\right], \tag{2}$$

where n is the number of apertures and B_1 and B_2 are the phase velocities in the coupled guides.

Assuming that $B_1 = B_2$, it is seen that the number of coupling slots, n, can be adjusted to give complete power transfer ($E_2 = 1$) for a particular c. However, c is a function of frequency and, as the frequency is increased, the number of slots will be insufficient to obtain complete power transfer, while at lower frequencies the number of coupling slots will be too large to obtain complete power transfer.

For the 4-kmc coupler, the two-polarization waveguide can be chosen to be square waveguide 1.790 inches on a side. The dominant mode cuts off at 3300 mc, and there are no modes with cutoffs in the 4-, 6- or 11-kmc bands. A coupling interval of four wavelengths was chosen to obtain complete power transfer from the square waveguide to a parallel rectangular waveguide ($n = 16$). The coupling obtained in each quarter wavelength interval must be equal[2] to $(\pi/2)/16 = 0.09830$. A slot 1.215 inches long in the sidewall of a rectangular waveguide 1.752 by 0.872 inches that couples to a 1.790-inch square waveguide will yield a coupling coefficient $c = 0.09830$ at 3870 mc. This calculation assumes that we have adjusted B_1 to be equal to B_2. (The method for doing this will be discussed later.) For 16 such slots, the amplitude of the wave emerging from port 3 of Fig. 4 when we excite port 1 is given by

$$E_1 = \cos nc\sqrt{\left(\frac{B_1 - B_2}{2c}\right)^2 + 1} - j\,\frac{B_1 - B_2}{2c}\,\frac{\sin\left[nc\sqrt{\left(\frac{B_1 - B_2}{2c}\right)^2 + 1}\right]}{\sqrt{\left(\frac{B_1 - B_2}{2c}\right)^2 + 1}}. \tag{3}$$

For the slot 1.215 inches long, we find at 3700 mc:

$$\begin{aligned} c &= 0.1203, \\ E_2 &= 0.940, \\ E_1 &= 0.342. \end{aligned}$$

At 4200 mc:

$$\begin{aligned} c &= 0.0755, \\ E_2 &= 0.940, \\ E_1 &= 0.355. \end{aligned}$$

Translating E_1 into uncoupled energy, we get the curve of Fig. 5, which shows that the insertion loss straight through the square waveguide (port 1 to port 3 of Fig. 4) is only about 10 db at the band edges.

These -10-db signals pass through the second coupler for the other polarization of 4 kmc with small loss and are completely reflected by the taper to 6-kmc dominant-mode two-polarization waveguide. The reflected signal suffers another 10-db loss in returning through the first coupler for 4 kmc, and emerges as a reflected signal that is only 20 db down. This is far from the objective of 30- to 35-db return loss. The above analysis leads to abandonment of the simple slot as a suitable element in the construction of directional couplers for systems combining networks.

A method has been found to modify the shape of the longitudinal

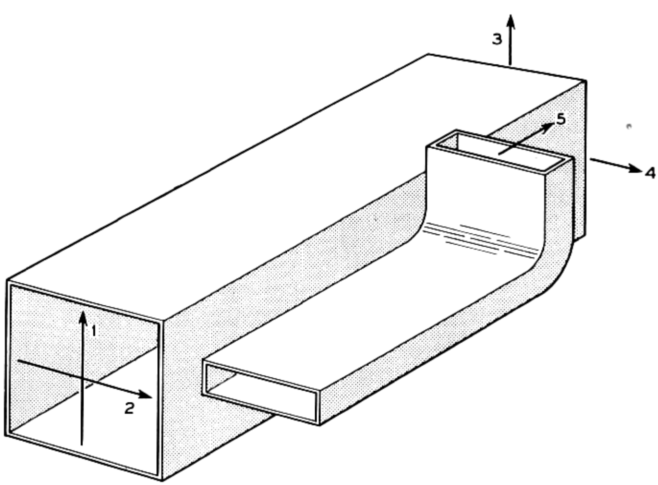


Fig. 4 — Accessible terminal pairs for directional couplers used in systems combining networks.

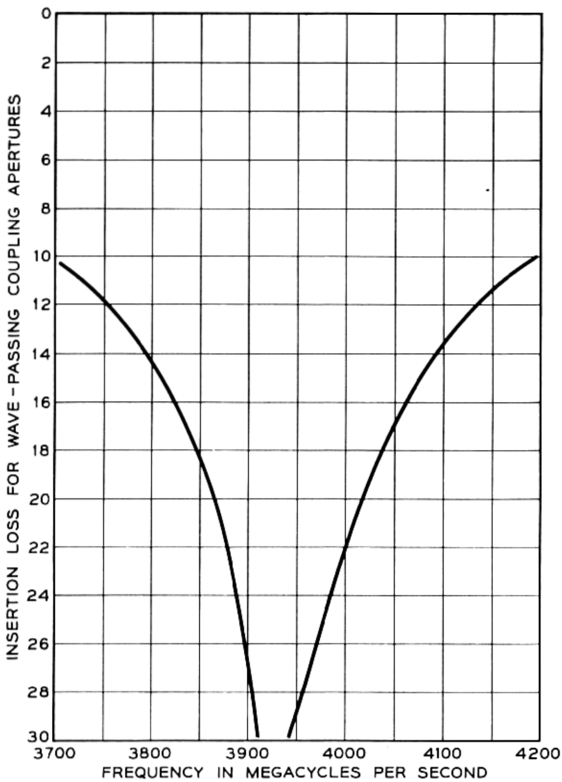


Fig. 5 — Insertion loss for wave passing coupling apertures.

slot to flatten the coupling-versus-frequency characteristic and reduce the amount of uncoupled energy. This structure is shown on Fig. 6. Placing the short piece of wire in the slot results in the coupling-loss-versus-frequency characteristic shown in Fig. 3. The wire is made approximately a quarter-wavelength long at 6 kmc, and results in a rejection peak at this frequency. The trial of a network element consisting of a wire resonator in the slot was suggested by viewing the coupling slot by itself as a short section of waveguide extending through the coupling plate. A resonant element placed in this approximately rectangular section of very short waveguide should be able to stop transmission at a particular desired frequency.

The frequency of the rejection peak is controlled by the total length of the wire while the width of the rejection peak is controlled by the location of the point at which the wire is bent and attached to the side

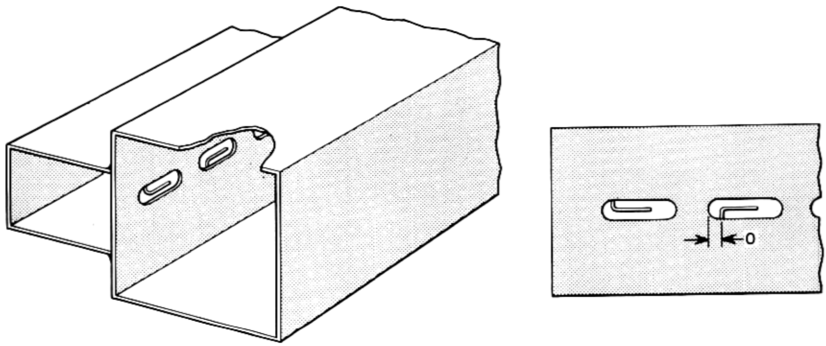
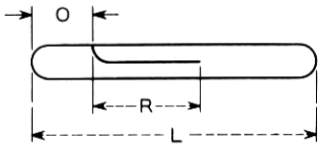


Fig. 6 — Directional coupler structure with modified slots.

of the slot (dimension O). Moving the point of attachment of the wire toward the center of the slot widens the rejection peak, but the left end of the curve, f_c, is determined by the waveguide size. It is seen that, as the width of the rejection peak is varied (wire offset changed), the slope of the coupling across the 4-kmc band can be changed. The width of the rejection peak is adjusted to produce a minimum of coupling near the center of the 4-kmc band. For each frequency in the lower part of the band, there will then be a frequency with the same coupling amplitude in the upper part of the band. With a series of such coupling elements, correct coupling amplitude for complete power transfer can be obtained at two frequencies in the 4-kmc band. Fig. 7 gives some data on the relation between resonator offset and coupling characteristics. The uncoupled energy at the two ends and the center of the 4-kmc band can be reduced to about 20 db, resulting in a return loss limitation of about 40 db due to the uncoupled energy.

The phase velocities in both the square waveguide and the rectangular waveguide are modified by the presence of the coupling slots. The amplitude and phase of the wave scattered in the forward direction in each waveguide can be calculated from (1). This scattered wave is added to the incident wave to obtain the phase shift caused by each coupling aperture. In the Appendix it is shown that an aperture-loaded rectangular 1.752- by 0.872-inch waveguide will produce the desired phase-velocity match with a square 1.790-inch waveguide near the center of the 4-kmc band.

The measured characteristics of a slot with resonator are shown in Fig. 8. The dimensions of the slot, resonator and offset have been experimentally adjusted to give the same coupling at the two edges of the



CURVE	DIMENSIONS		
	O	R	L
A	0.193	0.396	0.976
B	0.195	0.396	0.978
C	0.199	0.396	0.982

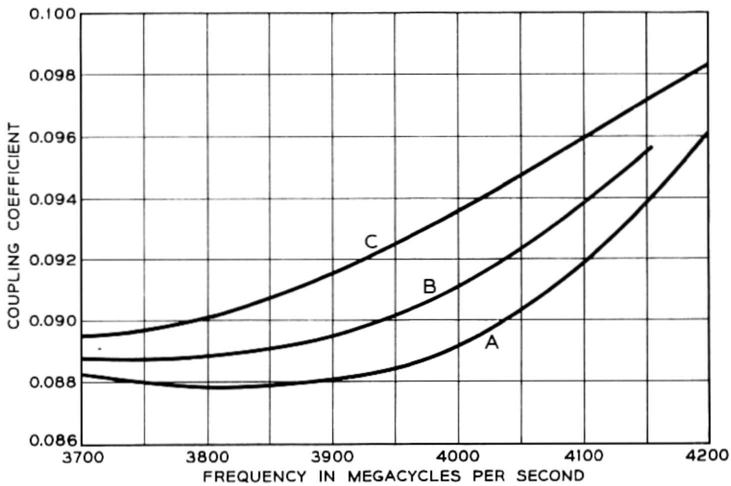


Fig. 7 — Effect of changing offset on coupling in 4-kmc band.

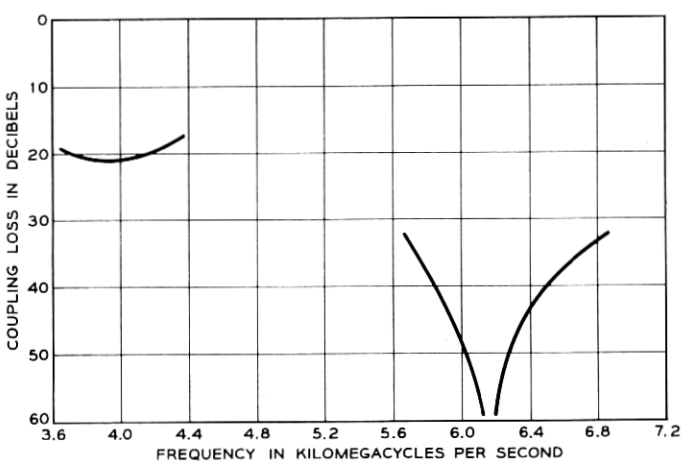


Fig. 8 — Measured response of single slot with resonator.

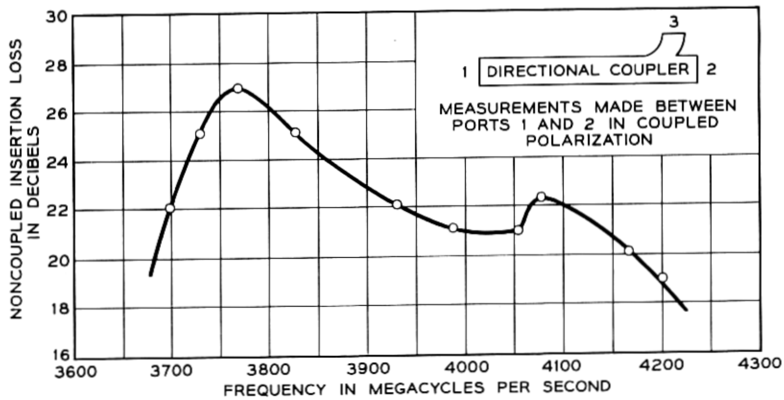


Fig. 9 — Straight-through loss for TD-2 band coupler.

4-kmc band, and a combination of 16 such slots has been made to give the coupling required for complete power transfer ($\sin nc = \pi/2$) at two points in the band.

The effect of the slot loading on the phase velocities in the square and rectangular waveguides is such that the phase velocities in the two waveguides can be exactly matched at only one frequency (approximately midband). At the two frequencies where the coupling is correct for complete power transfer, the phase velocity mismatch will limit the power transfer, so that the straight-through loss from port 3 to port 1 in Fig. 4 is only about 26 db. Fig. 9 shows the straight-through loss for a coupler with 16 coupling elements of the type shown in Fig. 8. The variation in coupling loss with frequency and the mismatch of phase velocity in the two waveguides both contribute to produce a straight-through loss of 19.5 db at the band edges and 21 db at the band center. The straight-through loss is now high enough to enable the realization of 30- to 35-db return loss on production models when other small reflections are included.

The resonant wires not only flatten the coupling characteristic in the 4-kmc band; they also reduce the coupling coefficients of the slots over the 6- and 11-kmc bands to such a small amount that very little energy (a few tenths of a db) is lost when these frequencies pass through the 4-kmc couplers. The only remaining deficiency of the network now is the return-loss performance at 11 kmc.

The slots in the 4-kmc coupler were spaced 1.440 inches apart, which is $\frac{5}{4}$ guide wavelengths at 10,740 mc, so the reflections should cancel at this frequency. However, the return-loss characteristic reached almost 23 db near 11 kmc. This was increased to over 30 db by putting in another

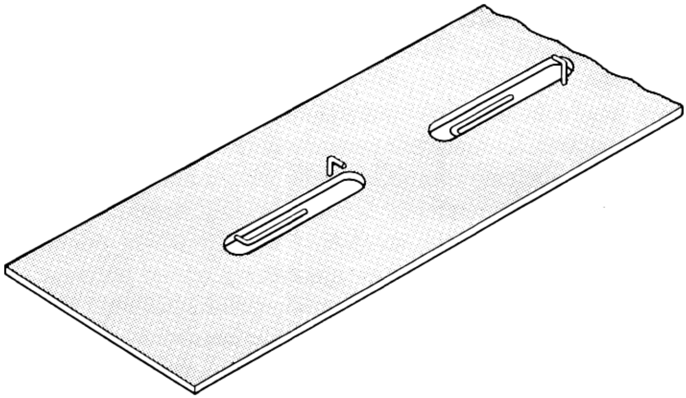


Fig. 10 — Coupling plate with added set of 11-kmc resonators.

set of wire resonators tuned to 11 kmc. There was insufficient room to place both the 6-kmc and 11-kmc resonators entirely within the slot, so the 11-kmc resonator was soldered in perpendicular to the plane of the coupling plate and it projected into the rectangular guide as shown in Fig. 10.

A similar design was produced for the 6-kmc directional coupler to be used in the systems combining network. In the 6-kmc version, it was found necessary to place two resonant wires in each slot, as shown in Fig. 11, in order to obtain adequate suppression of the 11-kmc energy. There was ample room for this, since the slot lengths in the 6-kmc directional coupler are considerably greater than a half wavelength at 11

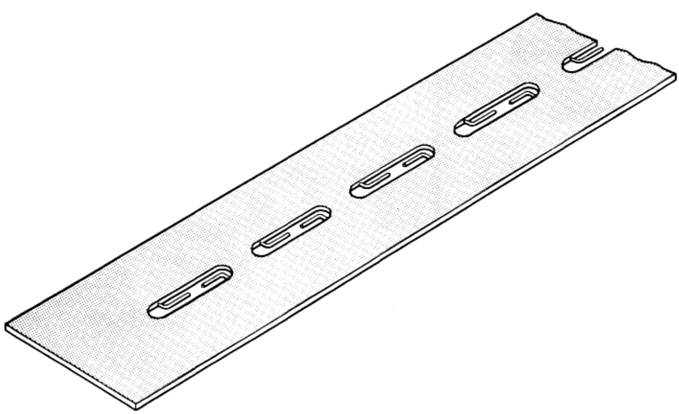


Fig. 11 — Coupling plate for 6-kmc coupler.

kmc. The energy transfer in the 6-kmc band was complete enough, so the straight-through loss exceeded 22 db over the band and the return loss due to this factor was 44 db.

Separation of the 11-kmc polarizations is accomplished by means of a polarization separation network previously described by Ohm.[6]

The complete systems combining network for both polarizations of all three common carrier frequency bands has eight accessible ports and 36 independent elements in the scattering matrix, which defines network performance. Fig. 1 illustrates the numbering of the ports, with port 1 and port 2 being the two dominant-mode polarizations in the square or circular waveguide. Table I summarizes the measurements made of the magnitude of the most important scattering elements in the three frequency bands of interest. The loss in the various desired signal paths is seen to be about $\frac{1}{2}$ db, and the impedance match at all ports is seen to exceed 30-db return loss. The cross-couplings in the various undesired signal paths are seen to range from 18 db to higher losses.

A photograph of a laboratory model of the systems combining network

Table I — Magnitude of Scattering Coefficients for Complete Systems Combining Network (Worst Performance at Any Point in the Band)

Element (see Fig. 1)	Measurement	Decibels Loss at		
		4 kmc	6 kmc	11 kmc
S11	Return Loss	33	30	30
S33	Return Loss	35	—	—
S55	Return Loss	*	38	—
S77	Return Loss	*	*	30
S88	Return Loss	*	*	30
S13	Insertion Loss	0.4	20	28
S14	Insertion Loss	40	50	40
S15	Insertion Loss	*	0.6	18
S16	Insertion Loss	*	33	40
S17	Insertion Loss	*	*	0.5
S18	Insertion Loss	*	*	40
S25	Insertion Loss	*	0.5	18
S34	Insertion Loss	50	—	—
S35	Insertion Loss	*	45	—
S36	Insertion Loss	*	55	—
S37	Insertion Loss	*	*	40
S38	Insertion Loss	*	*	30
S56	Insertion Loss	*	50	—
S57	Insertion Loss	*	*	35
S58	Insertion Loss	*	*	36
S78	Insertion Loss	*	*	50

* means the guide is beyond cutoff for this frequency.
— means the coefficient is of no interest at this frequency.

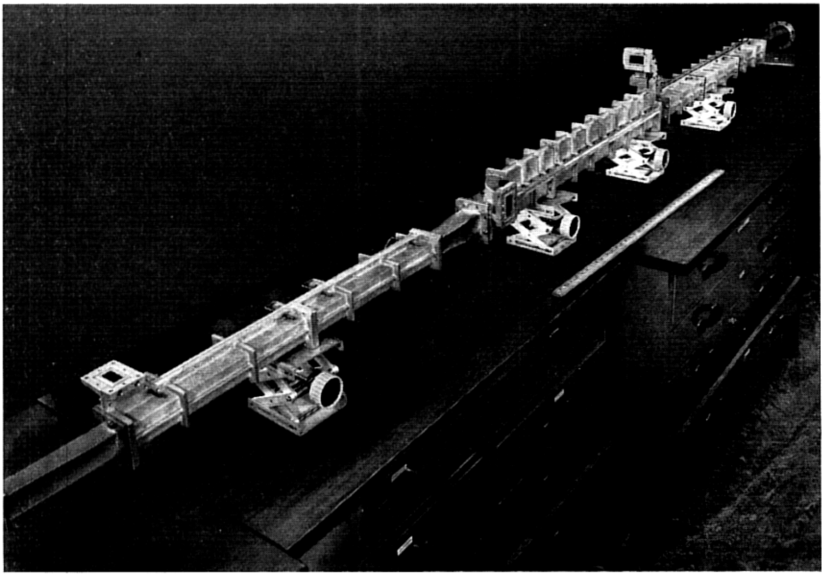


Fig. 12 — Laboratory model of systems combining network.

for separating polarizations of the 4-kmc and 6-kmc bands is shown in Fig. 12. The total length for combining both polarizations of all three frequency bands is 15 feet $10\frac{1}{2}$ inches.

There will be many repeater installations where all three frequency bands in both polarizations will not be used. In these cases, less than a full complement of directional couplers can be used. It is expected that there will be a large number of stations at which the horn reflector antennas will be fed with one polarization of 4 kmc and two polarizations of 6 kmc. The systems combining networks then need consist of only a circular-to-square taper, one 4-kmc directional coupler, a taper to 6-kmc dominant-mode square waveguide, one 6-kmc directional coupler and a transition from square to rectangular guide for the remaining 6-kmc polarization. The over-all length for this version of the systems combining network is less than 9 feet. In the case where only one polarization of each band is used, the total length would be 9 feet 1 inch.

CONCLUSION

A method has been developed for designing a systems combining network which separates by polarization and frequency the signals received by a common antenna for the 4000-mc TD-2, the 6000-mc TH and the 11,000-mc TJ radio relay systems. The design results in a theoretical reflected-signal limitation of about 40 db, while production models realize well over 30 db return loss. Insertion loss in the desired signal paths is about $\frac{1}{2}$ db, while couplings between various other ports range from 18 db to over 50 db. The structure is mechanically simple and can be fabricated to the required tolerances by conventional machining techniques. The same principles can be utilized for other frequency bands. However, new designs must be based on experimentally determined characteristics for the slots and resonators.

ACKNOWLEDGMENTS

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APPENDIX

The amplitude of the wave coupled between two guides has been computed by means of (1) for the coupling coefficient

$$c = M \frac{\pi}{2} \sqrt{\frac{\lambda_{g1} \lambda_{g2}}{a_1{}^3 b_1 a_2{}^3 b_2}}. \tag{1}$$

The amplitude of the wave scattered forward in guide 1 due to a wave incident in guide 1 can be calculated from the equation

$$c_{11} = M \frac{\pi}{2} \sqrt{\frac{\lambda_{g1} \lambda_{g1}}{a_1{}^3 b_1 a_1{}^3 b_1}} = M \frac{\pi}{2} \frac{\lambda_{g1}}{a_1{}^3 b_1},$$

and, of course, for guide 2,

$$c_{22} = M \frac{\pi}{2} \frac{\lambda_{g2}}{a_2{}^3 b_2}.$$

With c in coupling per unit length, the corrected phase shift per unit length is given by

$$\beta_{11}' = c_{11} + \beta_{11},$$

where β_{11}' is the corrected phase shift per unit length, and β_{11} is the unloaded guide phase shift per unit length. At 3900 mc, we have set $c = 0.0682$ per inch. Then, with $a_1 = b_1 = 1.790$ inches, $a_2 = 1.752$ inches and $b_2 = 0.872$ inch, we have

$$\beta_{11}' = 1.158 + 0.0682 \sqrt{\left(\frac{1.752}{1.790}\right)^3 \left(\frac{0.872}{1.790}\right) \left(\frac{0.5227}{0.5508}\right)}$$

$$= 1.158 + 0.045$$

$$= 1.203 \text{ radians per inch,}$$

$$\beta_{22}' = 1.099 + 0.0682 \sqrt{\left(\frac{1.790}{1.752}\right)^3 \left(\frac{1.790}{0.872}\right) \left(\frac{0.5508}{0.5227}\right)}$$

$$= 1.099 + 0.104$$

$$= 1.203 \text{ radians per inch.}$$

The 1.790-inch square guide and the 1.752- by 0.872-inch rectangular guide will now have the same phase velocity at 3900 mc, when they are loaded by slots that have a coupling coefficient equal to 0.0682 per inch.

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