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Channel combiners for radio-frequency transmitters

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SUMMARY
The essential requirements of channel combiners for broadcasting and communication circuits are listed. Circuit diagrams and descriptions of many possible types of combiners are given.

1 Introduction

To satisfy the needs of broadcasters or those who wish to communicate by radio telephone it is often necessary to transmit several frequencies from the same site. When some of these frequencies are closely spaced and are required to have approximately the same ground coverage it is desirable to transmit them from the same aerial. This requires the use of a channel combiner.

There are many advantages in using a single aerial with a channel combiner. Amongst these, the option to share aerials minimizes the proliferation of masts and towers, and enables less obtrusive and cheaper structures to be used. Transmitting from the same aerial and the same height enables similar ground coverage to be obtained and the channel combiner goes some way towards reducing the inevitable and undesirable interaction between aerials in close proximity to each other. With high-power v.h.f. or u.h.f. broadcasting systems it is often necessary to employ aerial elements on all sides of the mast to give all-round coverage. Also, at v.h.f. or u.h.f., several tiers of elements may have to be used to provide aerial gain and so keep transmitter power and running costs down to a reasonable level. (For example, a typical high-power transmitting aerial for v.h.f./f.m. sound broadcasts, capable of radiating up to five 40 kW transmissions, completely occupies 24 m of vertical mast space. Clearly it would be extremely costly to make separate equivalent aerial provision for each transmission.) For radio-telephone work and at the low-power end of the scale, where aerials are usually much simpler, the economics of combining are not so clear-cut but in many situations it is still structurally or environmentally highly desirable to combine.

Channel combiners may be produced in many different configurations, depending on the frequency of operation or on the input power levels; this paper describes the essential properties of most combiners and gives circuit diagrams of many different types that are commonly used today. A description of combiners in use in 1962 was published by Shone and Wharton.[1]

It is interesting to note that reversed combiners are often used to split signals from a single aerial between a number of different receivers tuned to different frequencies. Circuits identical to those of channel combiners can be used but there is no requirement to carry high power and low insertion losses are not usually considered to be so important.

2 Essential Requirements

The purpose of a channel combiner is to enable several transmissions on different frequencies to be radiated simultaneously from a single aerial system. The essential requirements of such a combiner are that:

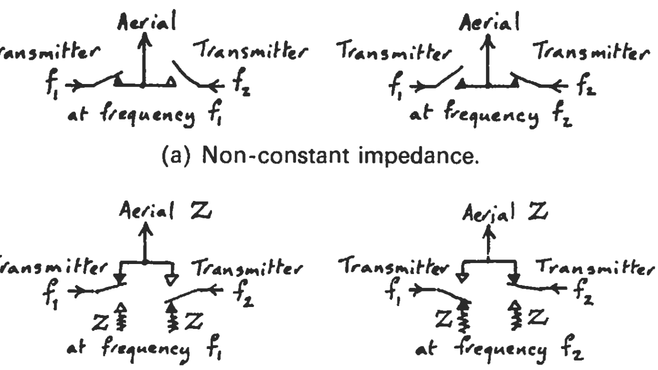
(1) It must be capable of accepting the required power of each transmitter and transferring it to the aerial with a minimum of loss. It must, of course, also present a reasonably well-matched impedance to each transmitter (i.e. its voltage reflection coefficient must not add significantly to that of the aerial).

(2) It must provide high cross-losses (high-attenuation paths) between transmitters on different frequencies. The reasons for this are two-fold:

First, the isolation should generally be greater than 30 dB in order to ensure safety. Secondly, the coupling must be kept low to avoid the creation of intermodulation frequencies which may interfere with other services. (See Sect. 4.)

(3) It must provide a distortion-free path for all input frequencies and their sidebands, i.e. the attenuation across an input channel should be reasonably constant and the phase change across the channel should be almost proportional to frequency. This implies that the group delay must be almost constant with frequency.

It is advantageous if the combiner can present a high-loss path between transmitters and the aerial at any intermodulation frequency. It is also advantageous, with some transmitters, if the combiner can present a 'constant impedance' (i.e. a low reflection co-efficient) at each input



(a) Non-constant impedance.

(b) Constant impedance.

Fig. 1. Combiners viewed as 'switches'.

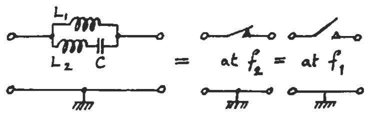


Fig. 2. Lumped component 'frequency switch'.

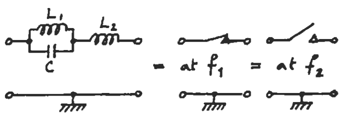


Fig. 3. Lumped component 'frequency switch'.

port over a very wide band of frequencies. This can improve transmitter stability and help to attenuate spurious frequencies.

3 General Principles of Operation

Most combiners include some frequency-dependent component or group of components which effectively behave as an open switch at one frequency and as a closed switch at another frequency as shown in Fig. 1. (Some exceptions to this rule are shown in the Appendix.)

At low-frequencies (l.f.) 30–300 kHz, medium-frequencies (m.f.) 0·3–3 MHz, and high-frequencies (h.f.) 3–30 MHz rejector and acceptor circuits comprising lumped components are used as 'frequency switches'. At very-high-frequencies (v.h.f.) 30–300 MHz, ultra-high-frequencies (u.h.f.) 300–3000 MHz and super-high-frequencies (s.h.f.) 3–30 GHz the frequency-dependent components are long open-circuit or short-circuit transmission lines, resonators, or parallel transmission line paths which differ in length.

3.1 Lumped Components

Figure 2 behaves as a closed switch when

$$f_2 = \frac{1}{2\pi\sqrt{L_2 C}}$$

and as an open switch at a lower frequency f_1, when

$$f_1 = \frac{1}{2\pi\sqrt{(L_1 + L_2)C}}$$

Figure 3 behaves as a closed switch when

$$f_1 = \frac{1}{2\pi}\sqrt{\frac{L_1 + L_2}{L_1 L_2 C}}$$

and as an open switch at a higher frequency f_2, when

$$f_2 = \frac{1}{2\pi\sqrt{L_1 C}}$$

3.2 Long Transmission Lines

In the rest of this paper 'transmission line' will be taken to mean open-wire line, coaxial line or strip line, whichever is appropriate to the frequency and power under consideration. Wavelength means wavelength in the transmission line and therefore takes into account the velocity factor of lines which have dielectric insulation.

If a short-circuit transmission line stub is connected across a line joining a generator G to a load R as shown in Fig. 4 the effect is to close a switch across the junction point at frequencies when the stub becomes an even number of quarter-wavelengths long and to open the switch when the stub becomes an odd number of quarter-wavelengths long.

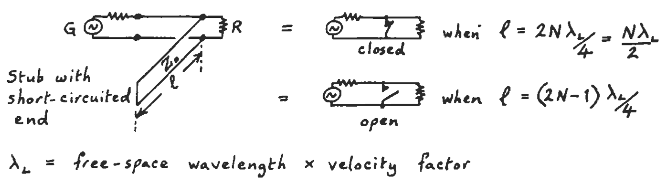


Fig. 4. Long-line 'frequency switch'.

If an open-circuited stub is used the switching effect is reversed.

The universal admittance of a stub is given by

$$Y = \frac{1}{Z_0} \tan\left(\frac{360° l}{\lambda_L}\right) \quad \text{open-circuit end}$$

and

$$Y = \frac{1}{Z_0} \cot\left(\frac{360° l}{\lambda_L}\right) \quad \text{short-circuit end}$$

Y changes from zero (open-switch) to infinity (closed switch) for a frequency difference such that $l = \lambda_L/4$ at the difference frequency. Clearly the stub has to be physically long if the frequency difference is small.

3.3 Loosely Coupled Transmission Lines and Resonators

A fast rate of change of admittance with frequency can be obtained differently, at least over part of the frequency band, by using a short stub (usually a quarter-wavelength long) and coupling into it 'loosely'. This may mean electrically coupling into the high voltage end of the stub via a small capacitance (Fig. 5) or magnetically coupling into the high current end of the stub by means of a tapping point or loop (Fig. 6). Both arrangements constitute resonators.

In coaxial form Fig. 6(a) becomes Fig. 6(b) and Fig. 5(a) becomes Fig. 5(b).

At first sight it may appear that one has everything to gain by using a loosely coupled transmission line stub in preference to a long-line stub. However, the nature of lines is such that in the 'closed-switch' condition of Fig. 5 or the 'open-switch' condition of Fig. 6 the currents and voltages in the stubs are multiplied to such a level that the losses

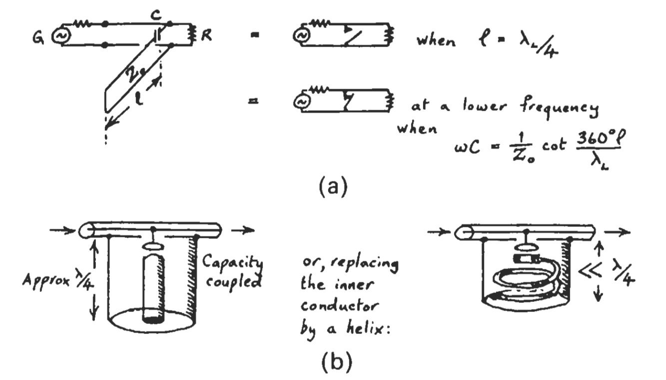


Fig. 5. Resonator 'frequency switch'.

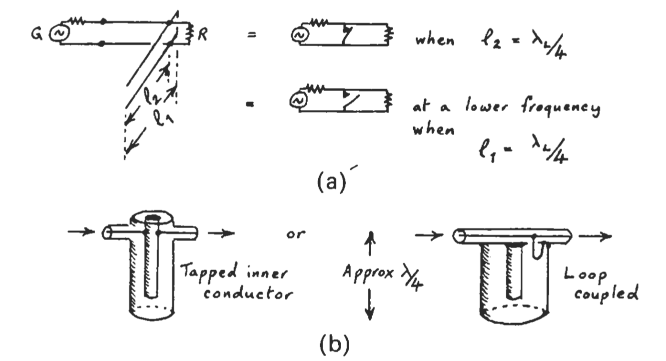


Fig. 6. Resonator 'frequency switch'.

are very nearly the same as those for a long line of the same diameter which changes condition at the same rate. So to ensure a low loss it is still necessary to employ a large diameter resonator. There are, however, considerable savings in 'copper' compared with using a long transmission line. There are even greater savings if a resonator with a helical inner conductor is used for low power at v.h.f. This is fully described in Ref. 7.

The voltage or current magnification follows a sine curve. Therefore if a line 90° long is magnetically tapped into at a point 10° from the short-circuit end, the voltage at the open-circuit end of the stub, V_{oc}, is equal to

$$\frac{\sin 90^\circ}{\sin 10^\circ} \times \text{applied voltage}$$

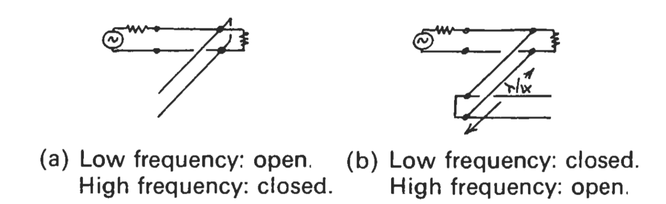
and the current at the short-circuit end of the stub, I_{sc}, is

$$\frac{V_{oc}}{Z_{0\,\text{stub}}}$$

The characteristic impedance of the stub does not have to be the same as that of the rest of the transmission system. Losses are least if it is about 70 ohms.

If it is required to reverse the frequency order of 'open-switch' and 'closed-switch' conditions for a given resonator system this may be carried out by connecting a quarter-wave of low-loss feeder between the through-line and the tapping point on the stub as shown in Fig. 7. Intermediate lengths of connecting feeder can be used to vary the 'open' and 'closed' bandwidths.

Fig. 7. Method of reversing frequency order.



(a) Low frequency: open. High frequency: closed.
(b) Low frequency: closed. High frequency: open.

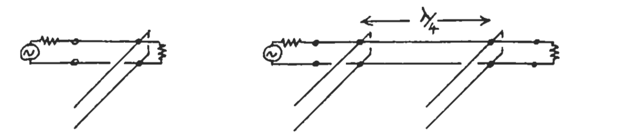
Various waveguide cavities can, of course, be used for u.h.f. or s.h.f. in place of coaxial cavities.

Care is needed in the design of a resonator to avoid voltage flashovers and hot-spots at high current points. The loss tangent of insulators must be low because, apart from pure efficiency considerations, hot insulators can significantly lower flashover voltages. Resonators must be dimensionally stable with temperature and, if necessary, constructed from low expansion alloys plated with copper or silver. If the equipment is to be used at high altitudes, due allowance must be made for reduced flashover voltages at the lower pressures.

3.4 The Effect of Loss in a Resonator and the Use of Multiple Resonators

The effect of loss in a resonator is to replace what would be an 'open-switch' by a very high resistor and what would be a 'closed-switch' by a very low resistor. Consequently through-losses in a combiner become non-zero and cross-losses become finite.

Improved attenuations for cross-losses, improved pass bandwidths and reflection coefficients can be obtained by using two or more resonators spaced a quarter-wavelength apart as shown in Fig. 8.



(a) Single resonator. (b) Double resonator.

Fig. 8. Single and double resonators.

Where a single resonator provides an attenuation greater than 3 dB, two resonators spaced a quarter-wavelength apart will give about double that attenuation. The resistive loss in the pass-band is also doubled when two resonators are used but at the edges of the pass-band, reflection losses are decreased. The rules are that reflection losses smaller than 1 dB (corresponding to a voltage reflection coefficient of 45%) are decreased when two resonators are used. The percentage improvement increases as reflection coefficients decrease. For example, 30% becomes 18%, 20% becomes 8%, and 10% becomes 2%.

4 Intermodulation

4.1 Frequencies

When two or more transmission frequencies f_a, f_b, etc., are combined into one aerial system, there is a possibility of intermodulation taking place due to finite coupling between transmitters. The intermodulation frequencies which are most likely to cause interference with other services are of the form $2f_a - f_b$ or $f_a + f_b - f_c$. This is because these frequencies remain in the same frequency band and are therefore radiated efficiently by the aerial system. They are also more difficult to filter out than those which occur close to the harmonic frequencies. The disposition of these intermodulation frequencies is shown diagrammatically in Fig. 9. Other products of the form

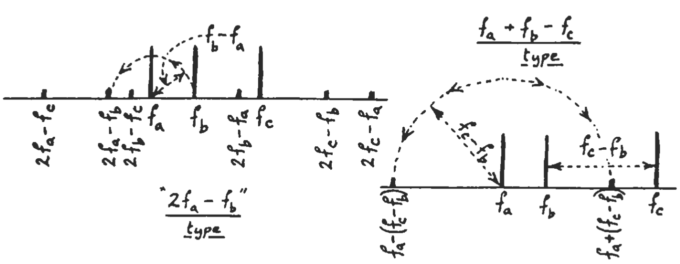


Fig. 9. Intermodulation frequencies.

$3f_a - 2f_b$ may be formed but often the level is too low to be measured. Figure 9 also shows how the expected intermodulation frequencies can be derived graphically.

4.2 Expected Levels

Intermodulation frequency $2f_a - f_b$ is generated by a low level of f_b getting through to the output stage of the f_a transmitter. There is then a conversion loss as f_b is converted to $2f_a - f_b$ which is dependent on the type of output stage of the transmitter and on the frequency separation of f_a and f_b. As an example, the conversion loss in a v.h.f. class C valved transmitter is 14 dB for a frequency separation of 2·2 MHz and 22 dB for a separation of 4·4 MHz. It is about 4 dB less for a solid-state transmitter.

The expected level of the radiated intermodulation product $2f_a - f_b$ can be obtained by adding together the following losses in dB:

(a) The cross-loss of f_b in getting to the f_a transmitter.
(b) The estimated conversion loss.
(c) The path loss of frequency $2f_a - f_b$ in getting from the f_a transmitter to the aerial.

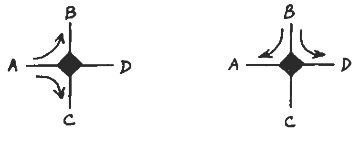
Intermodulation frequencies such as $f_a + f_b - f_c$ may be generated in any of the three transmitters by low levels of the other two frequencies reaching that transmitter. The expected levels can be calculated as shown above but no figures are available for the conversion losses.

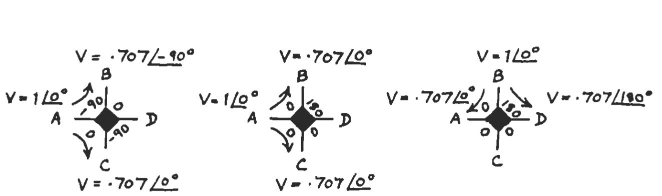
If radiated levels of intermodulation products are too high, one or more of the losses mentioned in (a) or (c) above must be increased. This is often done at v.h.f. by placing a notch filter in the output of say the f_a transmitter to reduce the level of f_b reaching it. It should be noted, however, that because f_b 'sees' the f_a transmitter as an unmatched load it is necessary to position the notch filter at a voltage maximum of f_b in order to ensure maximum effect.

5 Combiner Components: Hybrids and Circulators

Because radio frequency hybrids and circulators act as building blocks for a number of combiner circuits a brief description of them and their properties will be given here.

Fig. 10. Ideal hybrid power division.





(a) Quadrature type. (b) 0°–180° type.

Fig. 11. Phases of voltage vectors in hybrids

5.1 Radio Frequency Hybrids

A fuller description of hybrids is given in Ref. 2 but a summary will be given here.

An ideal hybrid as symbolized in Fig. 10 is a symmetrical four-port device in which power fed into any port, say A, is divided equally between two output ports B and C and no power is transferred to the opposite port D, provided that ports B and C are matched.

In the same way, power entering port B is split equally between ports A and D and no power is transferred to port C.

The hybrid comprises transmission lines, transformers or lumped components and is completely reciprocal and linear.

Two families of hybrids exist:

(a) the 90° (quadrature type) where the phases of the voltage vectors emerging from output ports differ by 90° and
(b) the 0°–180° type where the output voltage vectors are either co-phased or 180° out of phase, depending on which ports are used.

The two families may be symbolized as shown in Fig. 11, where the figures round the centre refer to the phase changes experienced in proceeding from one port to another.

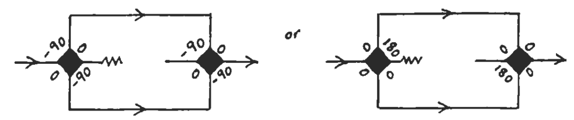
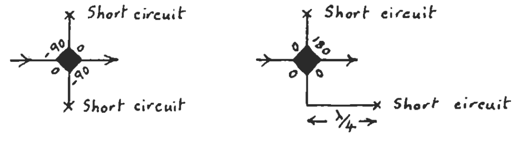


Fig. 12. Hybrids in back-to-back mode.

In practical combiner circuits hybrids either perform in back-to-back mode (Fig. 12), or in short-circuited-output mode (Fig. 13).

Many different forms of practical hybrids exist. The most common form of quadrature hybrid used in combiners is the '3 dB coupler' which is a coupled

Fig. 13. Hybrid in short-circuited output mode.



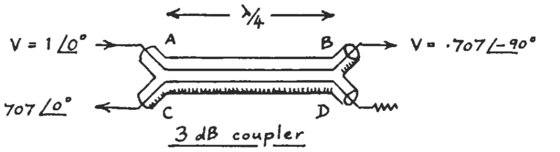


Fig. 14. 3 dB coupler hybrid.

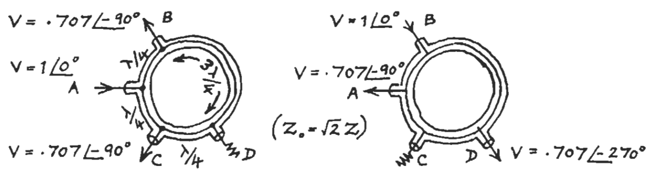


Fig. 15. Rat-race hybrid.

transmission line directional coupler specially designed to give approximately equal outputs at the design frequency (Fig. 14).

The amplitude ratio of its two outputs varies with frequency but the input impedance and the cross-loss and phase difference between outputs theoretically remain constant.

One type of 0–180° hybrid is the rat-race (Fig. 15), which is a $1\frac{1}{2}$-wavelength ring of transmission line or waveguide with a characteristic impedance equal to $\sqrt{2}$ times the input and output impedances.

The input impedance, output ratio and cross-losses all vary with frequency but reasonable operation can be obtained over a frequency bandwidth of about $\pm 6\%$. This bandwidth can, however, be improved to some extent by feeding ports A and C through $\frac{1}{2}$-wavelengths of $0{\cdot}84Z$ line and ports B and D through $\frac{1}{2}$-wavelengths of $1{\cdot}19Z$ line.

Another type of 0–180° hybrid is the split drum. Illustrated in Fig. 16, it is complicated to fabricate but has a good operating bandwidth and, theoretically, only its input impedance varies with frequency.

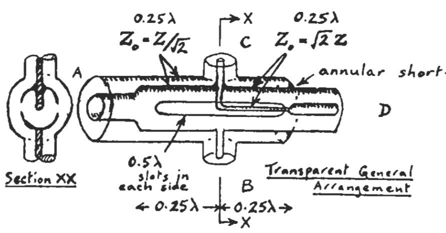


Fig. 16. Split-drum hybrid.

5.2 The Circulator

The circulator is a non-reciprocal transmission line or waveguide component normally having three ports. It contains a ferrite material and its operation is somewhat analogous to the mechanical behaviour of the gyroscope. (If one end of the axle of a gyroscope is supported and the other is free, a downward push on the free end may cause a movement of the free end to the left. An attempt to restore the position by pushing from the left will then result in a non-reciprocal downward movement.) Because it is a narrow-band device and has a fairly high loss, it only has a limited usefulness in channel combiners.

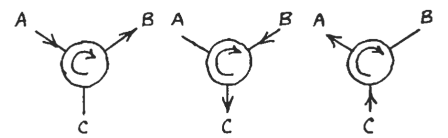


Fig. 17. Circulator power directions.

Shown in Fig. 17, the circulator requires an arrow to indicate the direction of 'power circulation'.

If port B is perfectly matched, then power entering port A will, apart from losses, be entirely transferred to port B and no power will be transferred to port C. Similarly, power entering port B will, apart from losses, be entirely transferred to port C provided that port C is perfectly matched. (If the output port is imperfectly matched the circulator has the remarkable property of depositing reflected power in a load attached to the third port. Therefore, as far as the generator is concerned, the output load appears to be perfect.)

6 Commonly Used Types of Combiner

Fifteen different types of combiners will now be described with comments on their design and the relative advantages and disadvantages of their use. Many of the designs are related to each other and it is often possible to make variations on established designs. Unless it is otherwise stated, all combiners described here can be assumed to be of the constant impedance type (see Sect. 2).

Frequencies may be combined in parallel (Fig. 18) or, as more commonly used in broadcasting, in cascade (Fig. 19). In the parallel arrangement all units of the combiner have equal bandwidths and have equal power ratings (assuming equal transmitter powers).

In the cascade arrangement each unit, other than the first, has a narrow frequency band input and a wide frequency band input which is capable of accepting all frequencies which have already been combined up to that point. The unit closest to the aerial must of course have a higher power rating than units further back in the chain. Frequencies can be combined in cascade in any order, provided that units are suitably designed.

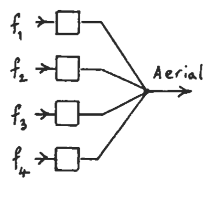
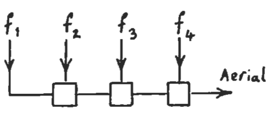


Fig. 18. Parallel combination.

Fig. 19. Cascade combination.



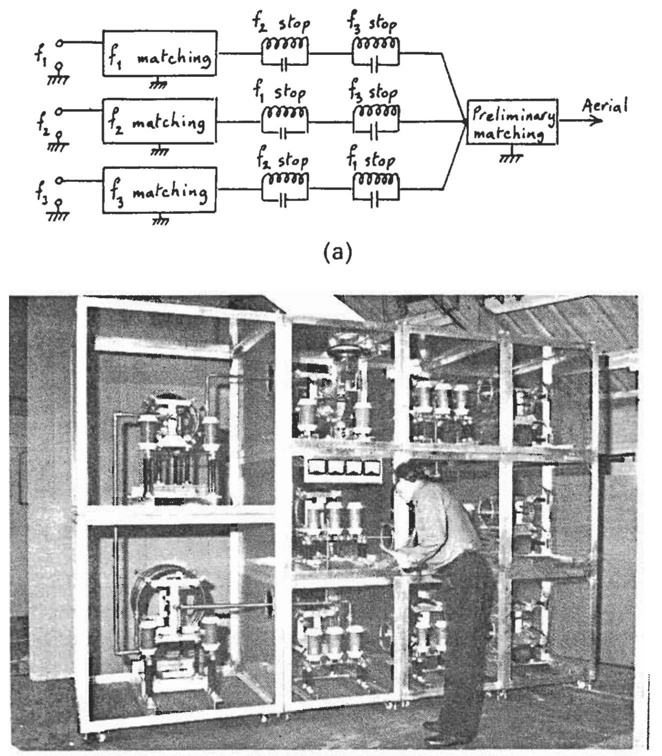


Fig. 20. (a) Lumped component non-constant impedance parallel combiner. (b) M.f. combiner for combining three 10 kW transmissions. *(C & S Antennas.)*

6.1 Non-constant-impedance Parallel Combiner using Lumped Components

This type of parallel combiner, shown in Fig. 20(a), is commonly used in m.f. and h.f. broadcasting. A practical arrangement is shown in Fig. 20(b). The components associated with each input frequency are usually housed in separate screened compartments to avoid coupling between inductors.

By installing preliminary matching components it may be possible to produce compromise aerial impedances at all input frequencies which minimize currents through the rejector circuits yet keep voltages down to a reasonable level. The individual-frequency matching components can be eliminated by building series components into rejector circuits but this makes the system much harder to adjust. Acceptor/rejector circuits of the types shown in Figs. 2 and 3 may be installed in shunt with any transmission path to increase the cross-loss between transmitters at the stop frequency. The power rating of components in rejector circuits must take into account both the stopped frequency and the pass-frequency. The pass-frequency usually provides the greatest currents which circulate in the rejector components. These are given by:

$$I_C = I_p[1-(f_s/f_p)^2]^{-1}$$

and

$$I_L = I_p[1-(f_p/f_s)^2]^{-1}$$

where I_C = current in the capacitor
I_L = current in the inductor
I_p = total resultant pass current
f_s = stop-frequency
f_p = pass-frequency

Note that component currents increase as the frequency ratio approaches unity.

The necessary voltage ratings of components are obtained by adding the peak r.f. voltages due to each frequency, taking amplitude modulation into account. The necessary current ratings are obtained by taking the root-mean-square of all carriers and sideband currents through the component. Narrow-band amplitude modulated m.f. or h.f. transmitters will normally operate quite satisfactorily into a non-constant-impedance combiner.

6.2 Non-constant-impedance Parallel Combiner using Resonators

This type of parallel combiner, shown in Fig. 21, could be used at v.h.f. or u.h.f. but is rarely used in broadcasting because the non-constant-impedance characteristic often leads to instability of the transmitters and the radiation of spurious frequencies.

The quarter-wave lines leading to the common-point output are a compromise quarter-wavelength at frequencies other than the pass-frequency. The pass-resonators can operate in identical pairs as shown in Fig. 21 or, to obtain greater cross-losses, they could appear in groups of three or four.

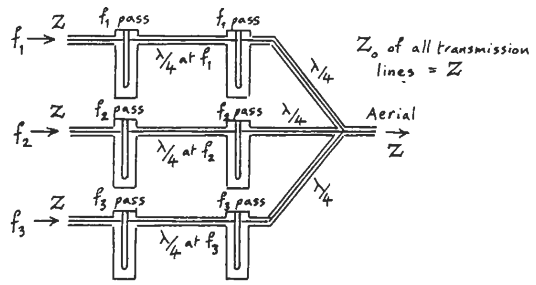


Fig. 21. Non-constant impedance parallel combiner using resonators.

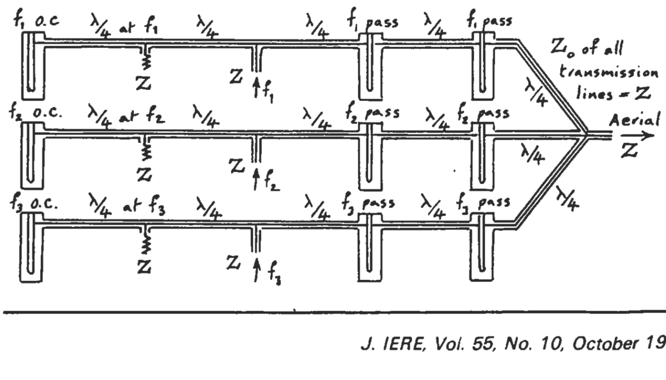
The advantage of using twos or fours is that there is mutual impedance compensation at the pass-frequency (see Sect. 3.4). The same sort of effect can also be obtained if three resonators are spaced a quarter-wave apart but in this case the centre resonator has to have 1·5 times the reactance slope of the other two resonators in order to compensate.

6.3 Constant Impedance Parallel Combiner using Resonators

The constant impedance version of the circuit given in Fig. 21 is shown in Fig. 22.

Most of the design remarks relating to Fig. 19 still apply but each transmitter now sees a reasonable input impedance over a wide band of frequencies. The resonators

Fig. 22. Constant impedance parallel combiner using resonators.



in each input chain present an open-circuit in shunt with the transmission line to which they are attached. The result of this is to open up the path to the aerial and to put an effective short-circuit across the balancing load at the input frequency. At other frequencies in the band the path to the aerial is closed and power from the transmitter is diverted to the balancing load. The characteristic impedance of all transmission lines is normally equal to the input and output impedances.

A disadvantage of this type of combiner is the difficulty of obtaining a good impedance match over each input frequency band if the number of input frequencies is greater than two.

6.4 Two-wavelength Ring (Maxwell Bridge)

This ring circuit (Fig. 23) is sometimes known as a Maxwell Bridge because, by suitable impedance manipulations and star-delta conversions, the two circuits can be shown to be identical at the design frequency. It is suitable for cascade combination.

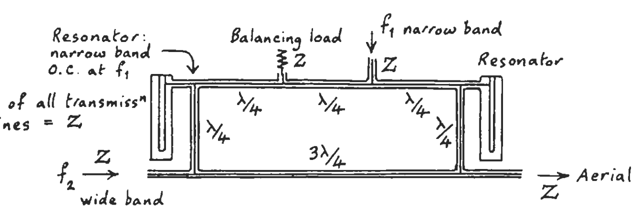


Fig. 23. Two-wavelength ring.

A true Maxwell Bridge combiner is described in the Appendix where it will be seen that conjugate impedance resonators have to be used and special arrangements have to be made for coaxial connections. Here, however, the resonators are identical and are both adjusted to present open-circuits in shunt with the transmission line ring at the narrow-band input frequency. The division of power from any input port can be deduced by comparing path lengths in traversing the ring by any route round the ring. For example, a signal entering the f_1 port can reach the aerial port via a $\frac{1}{2}$-wavelength path or via a $1\frac{1}{2}$-wavelength path; currents therefore reinforce and power is transferred. On the other hand, the same signal reaches the f_2 port via a $\frac{3}{4}$-wavelength path or via a $1\frac{1}{4}$-wavelength path; currents therefore cancel out and the two ports are isolated. Similar reasoning can be applied to other ports in the ring. At frequencies fairly well separated from f_1 which enter the f_2 port, the resonators behave as short-circuits or high susceptances across the ring. The cross-loss to the f_1 port is therefore reinforced and power is transferred to the aerial by virtue of the two stopped-off quarter wavelengths of line at the sides of the ring. The balancing load absorbs any frequencies other than f_1, which enter the f_1 port and also absorbs any frequencies which enter the f_2 port and manage to get past the resonators. The length of the ring should be electrically two-wavelengths at the f_1 frequency in order to obtain maximum cross-loss at that frequency.

It should be noted that the f_2 port is a wideband port. Its input characteristics are determined only by the two quarter-wave lines with high-susceptance resonators at their far ends and, because these are separated by $\frac{3}{4}$-wavelength, their reflection coefficients cancel to some extent.

Theoretical cross-loss bandwidths for a ring without resonators are as follows:

40 dB	$\pm 3{\cdot}3\%$ frequency
35 dB	$\pm 4{\cdot}4\%$ frequency
30 dB	$\pm 6{\cdot}0\%$ frequency

The cross-loss between transmitters for a frequency f_2, which is not far separated from frequency f_1, can often be improved by adjusting the complex impedance of the balancing load.

This type of combiner can also be used in a different mode by reversing the role of the input ports as shown in Fig. 24. The length of the ring is this time equal to two wavelengths at frequency f_2. There is a slight improvement in wide-band pass-loss and reflection coefficient when the ring is used in this mode.

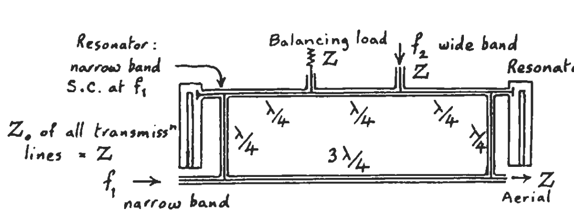
6.5 'Phillips Ring', 'Figure-8 Ring' (Maxwell Bridge)

Invented by Dr G. J. Phillips,[3] this ring circuit (Fig. 25), can also be derived from the Maxwell Bridge. It is suitable for cascade combination.

The resonators are identical and are usually adjusted to present open-circuits in shunt with the transmission line ring at the narrow band frequency. The division of power can be deduced by comparing path lengths in traversing from one port to another via different routes. For instance, when the resonators present open-circuits and are effectively removed from the ring, the shortest path from the f_1 input port to the f_2 input port is a half-wavelength. The simplest path by any other route is always one-wavelength. Hence vectors cancel and the two ports are isolated from each other.

At the narrow-band frequency f_1, the balancing load is similarly isolated from the input port and, since no voltage is expected at isolated ports, these ports can equally well be considered to be short-circuited. The power entering the f_1 port can then be considered to reach the aerial via a $\frac{3}{4}$-wavelength zig-zag path passing the two resonators on the way. Effectively short-circuited $\frac{1}{4}$-wave stubs are attached in shunt with the transmission line at each junction point. The sides of the ring should each be a $\frac{1}{4}$-wave long at the f_1 frequency.

Fig. 24. Two-wavelength ring.



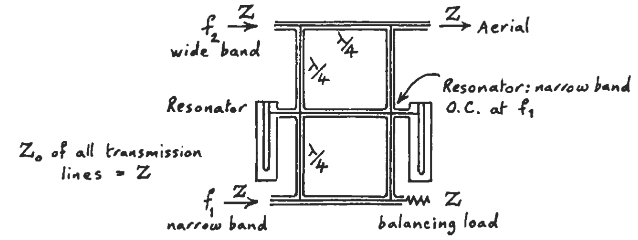


Fig. 25. Figure-8 ring.

At the f_2 frequencies, when the resonators present a short-circuit or a high susceptance in shunt with the transmission lines, the f_2 port is more effectively isolated from the f_1 port and power is transferred directly to the aerial via a $\frac{1}{4}$-wave line with effective $\frac{1}{4}$-wave stubs on either end. This gives a good wide-band characteristic because any imperfections of stub impedances tend to cancel out.

Theoretical cross-loss bandwidths for a ring without resonators are as follows:

40 dB	±2·7% frequency
35 dB	±3·6% frequency
30 dB	±4·7% frequency

The wide-band impedance, however, is better than that of the ring described in Section 6.4.

Once again, the cross-loss between transmitters for a frequency entering the f_2 port, which is not far removed from frequency f_1, can often be improved by adjusting the complex impedance of the balancing load.

6.6 Combiner for V.H.F. using Quadrature Hybrids and Resonators

The combiner shown in Fig. 26(a) uses 3 dB couplers (with built-in cross-overs for mechanical convenience) as quadrature hybrids. It is suitable for cascade combination. The resonators are identical and are adjusted to present short-circuits across the interconnecting transmission lines at the narrow-band input frequency f_1.

The right-hand hybrid then behaves in a short-circuited output mode at frequency f_1 as discussed in Section 5.1 and the resonators help to prevent frequency f_1 reaching the f_2 transmitter. Frequencies entering the f_2 port see the resonators as relatively high reactances in shunt with the interconnecting transmission lines and use the back-to-back hybrid mode to reach the aerial. If the resonators do not appear as complete open-circuits, some power from the f_2 transmitter will be reflected back into the balancing load instead of being transferred to the aerial. Power from the f_2 transmitter is prevented from reaching the f_1 transmitter only by the 'output balance' of the hybrids, which cannot be perfect at all frequencies in a system using 3 dB couplers.

Theoretical cross-loss bandwidths for rings without resonators are as follows, but it should be noted that these figures may be modified by aerial reflections:

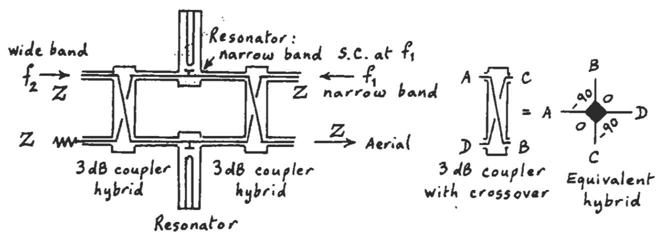
	3 dB coupler	2·97 dB coupler
40 dB	±9·0%	±12·6%
35 dB	±12·0%	±14·8%
30 dB	±16·0%	±18·0%

(Note the improved bandwidth if the coupler is over-coupled to 2·97 dB and the considerable improvement over the previous two combiners. A further increase in coupling reduces the 40 dB bandwidth but increases the 35 dB bandwidth.) Theoretically the input impedance should be perfect at all frequencies.

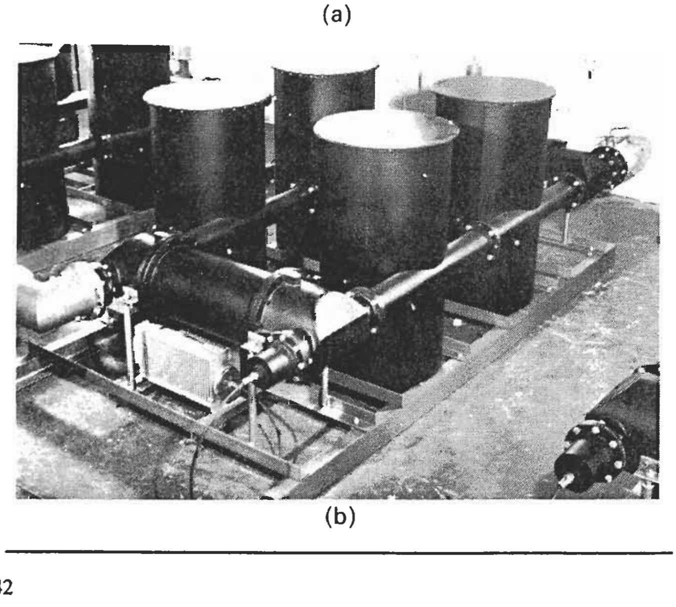
There is a considerable improvement in the cross-loss of f_1 to the f_2 transmitter if two resonators spaced by a $\frac{1}{4}$-wavelength are used in each interconnecting feeder. A practical v.h.f. arrangement of this is shown in Fig. 26(b). The wide-band through-loss is also considerably improved. All resonators can be identical but those closest to the f_1 transmitter need to have the highest power rating. The currents entering each of these resonators at the T-junctions at frequency f_1 are each equal to $\sqrt{2}$ times the current in a matched transmission line fed by the total f_1 input power. The effect of this is to give rise to twice the power loss that one would expect at first sight. Currents due to the f_1 frequencies are far lower and depend on the ratio of f_1 to f_2 and on the bandwidths of the resonators. The cross-loss between transmitters for a frequency entering the f_2 port, which is not far removed from frequency f_1, can often be improved by adjusting the complex impedance of the balancing load.

Through-loss and cross-loss parameters for a typical

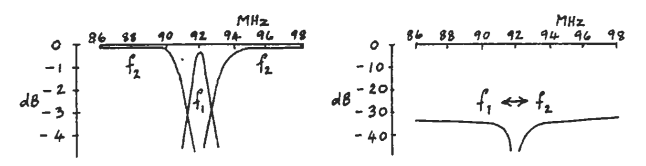
Fig. 26. (a) V.h.f. combiner using two 3 dB couplers and two resonators (b) Practical version using four resonators. The complete cascade combiner is designed to combine up to five 20 kW transmissions in the frequency band 88–108 MHz. *(Alan Dick and Co.).*



(a)



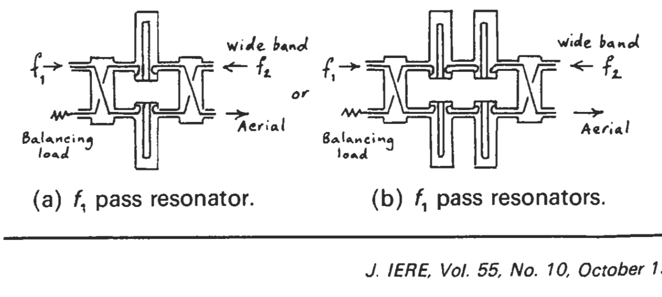
(b)



(a) Through-loss. (b) Cross-loss.

Fig. 27. Four-resonator v.h.f. combiner performance

Fig. 28. Combiner using band-pass resonators.



(a) f_1 pass resonator. (b) f_1 pass resonators.

four-resonator v.h.f. combiner operating in this mode are shown in Fig. 27.

It is also possible to use this type of combiner in a different mode whereby each interconnecting feeder contains a single loop-coupled resonator or pair of loop-coupled resonators to give a narrow-band pass in the back-to-back mode and a wide-band stop in the short-circuit output mode. The circuit would then be as shown in Fig. 28.

6.7 Combiner for U.H.F. Sound/Vision using 0–180° Hybrids and Resonators

The combiner shown in Fig. 29 is very similar in its mode of operation to that described in Section 6.6. However, because 0–180° split-drum hybrids (see Fig. 16) are used the resonators along the two interconnecting feeders have to be off-set by a $\frac{1}{4}$-wavelength to achieve the correct phase conditions.

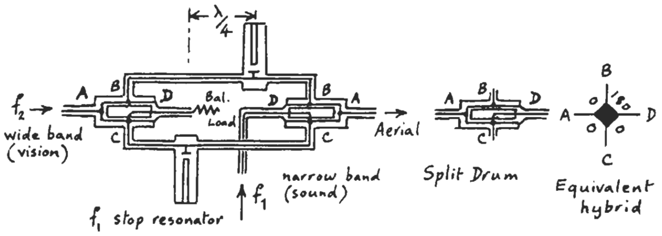


Fig. 29. U.h.f. combiner using hybrids and resonators.

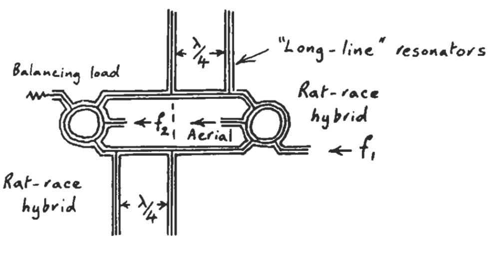
Theoretically the cross-loss, determined by the 'output balance' of the split-drum hybrids, should be perfect at all frequencies but the input impedance will vary with frequency. Extra resonators are often provided to shape the vision band.

6.8 Combiner for U.H.F. Channels using 0–180° Hybrids and Long Transmission Lines as Resonators

The frequency spacing of u.h.f. television channels in the UK is such that transmission line stubs of a reasonable length (see Sect. 3.2) can be used economically in back-to-back hybrid rings in place of resonators.

The first generation of BBC high-power u.h.f. combiners used rat-race rings (see Fig. 15) as hybrids as shown in Fig. 30. The two input bandwidths are approximately equal and the currents and voltages in the 'long-lines' do not rise above two times the matched-load input voltages. For two u.h.f. channels spaced typically 48 MHz apart the length of a 'long line' is about 1·5 m. Ends are normally open-circuits to give ease of adjustment. The length is such that they present good short-circuits across the inter-

Fig. 30. U.h.f. combiner using hybrids and long transmission lines.



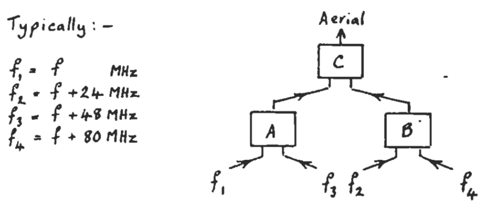


Fig. 31. Method of combining four u.h.f. channels.

connecting feeders at f_1 and reasonable open-circuits at f_2. Where four channels have to be combined it is usual to combine alternate pairs in two-channel combiners and then to make use of the periodic characteristic of a 'long-line' to combine the two pairs as shown in Fig. 31.

The 'long lines' on combiners A and B are all about 1·5 m long whilst those on combiner C are about 3 m long. The latter are adjusted to present short-circuits across the interconnecting feeders in channels f_1 and f_3 and reasonable open-circuits at channels f_2 and f_4 or vice-versa.

It can be seen that the periodic impedance characteristic of a 'long-line' is an advantage as far as 3rd-order intermodulation products are concerned because these products are naturally diverted to the balancing loads.

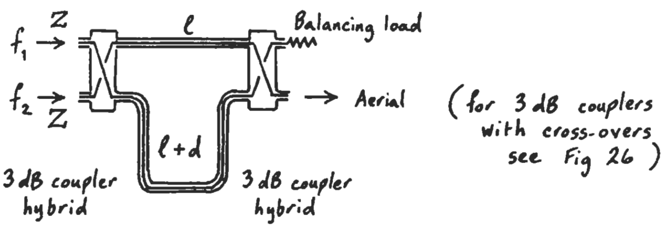
6.9 Combiner for Low-power U.H.F. Channels using Hybrids and Unequal-length Lines

The combiner shown in Fig. 32 uses 3 dB coupler quadrature hybrids in back-to-back mode with unequal-length interconnecting feeders. (For 3 dB couplers with cross-overs see Fig. 26(a).) The difference in length, d, is arranged to be an even number of $\frac{1}{2}$-wavelengths at frequency f_1 and an odd number of $\frac{1}{2}$-wavelengths at frequency f_2.

For a typical channel spacing of 48 MHz and polythene-dielectric feeder (velocity factor 0·67) d is about 2 m. As in Section 6.8 a typical four-channel combiner is formed by using a difference in feeder length of about 4 m and interleaving the channels. Many channel frequency spacings which are used in practice are not perfectly suited by any particular length-difference. A compromise then has to be made. As in Section 6.8 the 3rd-order intermodulation products tend to be absorbed in the balancing load.

A particular advantage of this type of combiner, which has no sort of resonator, is that it has a constant group delay (linear phase shift with frequency) across each input band. It can be considered that the whole circuit behaves as a frequency-dependent component.

Fig. 32. Low-power combiner using hybrids and unequal-length lines.



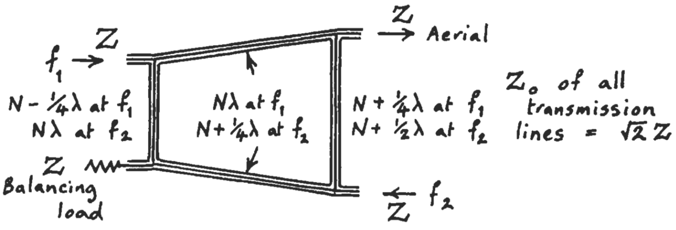


Fig. 33. Commutating line diplexer.

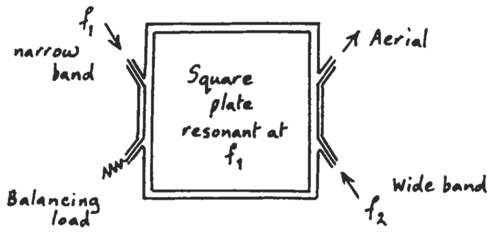


Fig. 35. Resonant plate combiner.

6.10 Commutating Line Diplexer for Combining U.H.F. Channels

A form of combiner, proposed by Mr A. B. Shone (formerly BBC), which as far as is known has not been used in practice, is shown in Fig. 33.

It consists of a ring of transmission lines whose lengths change from odd to even numbers of $\frac{1}{4}$-wavelengths from one channel to the next. With most practical channel spacings the lengths of lines have to be somewhat of a compromise to approach the commutating requirements, but it can be made to work sufficiently well in theory. The mean length of a side of the ring is a $\frac{1}{4}$-wavelength at the difference frequency of f_1 and f_2. The principle of operation can be seen by measuring path differences between different routes round the ring as described in Section 6.4. At first sight the transmission lines in the ring appear to be able to have any arbitrary characteristic impedance but it may be shown that a higher impedance than Z gives improved characteristics and that $(\sqrt{2})\ Z$ is about as high as should be used to avoid problems with standing waves.

This circuit also has the advantage that it has a constant group delay across each input band. Its lack of use probably stems from the difficulty of making adjustments during the development stage.

6.11 Travelling-wave Combiner for V.H.F. or U.H.F.

The travelling-wave combiner[4] uses a loop of feeder, which is one or two wavelengths long, as a resonator. The circuit is shown in Fig. 34 and, since it has a narrow-band input port and a wide-band input port, it is suitable for cascade combining.

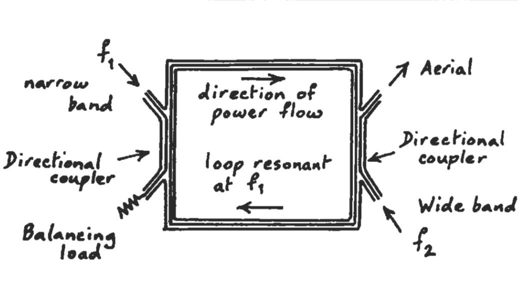


Fig. 34. Travelling-wave combiner.

The bandwidth required on the narrow-band port determines the number of wavelengths that can be used in the ring and the degree of coupling of the directional couplers. The degree of coupling is usually small, i.e. less than -6 dB, and to operate well the ring must have a low loss and be accurately matched at the narrow-band frequency. It should be noted that, with -6 dB coupling and a low-loss feeder, the amount of f_1 power circulating in the ring is almost four times the f_1 power entering the ring. The principle is similar to that described in Appendix 4 of Ref. 2. Power at frequency f_2 fails to excite the ring and is delivered straight to the aerial.

6.12 Resonant Plate Combiner for V.H.F. or U.H.F.

The resonant plate combiner has been used by Marconi Communication Systems to combine v.h.f. channels. The circuit is shown in Fig. 35. A full description is given in Ref. 5.

In principle it does not differ a great deal from the travelling-wave combiner but in this combiner a flat metal plate, which is electrically isolated from the sides of an enclosure, is made to resonate at the f_1 frequency by means of a directional coupler. A further directional coupler extracts power for the aerial.

6.13 The 'Rotamode' Combiner for U.H.F.

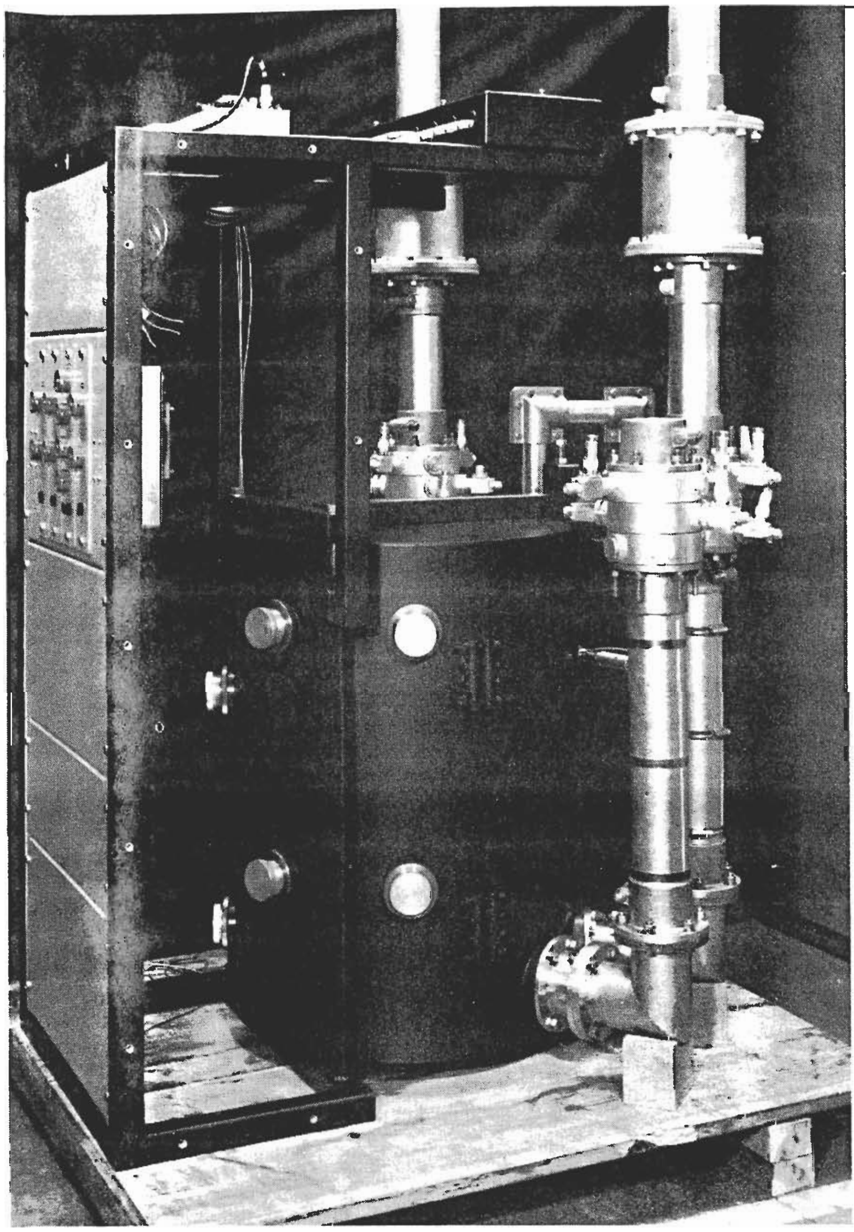
Another type of combiner which operates very similarly to the travelling-wave combiner is the 'Rotamode' designed by Mr R. Hutchinson of Marconi Communication Systems. A full description of this is given in Ref. 5 and it is illustrated in Fig. 36(a). It consists basically of a pair of directional couplers set in the sides of a cylindrical waveguide cavity which will support a TE_{111} mode as shown in Fig. 36(b). The cavity may be a $\frac{1}{2}$-wavelength long or one-wavelength long.

The cavity is designed to resonate at the narrow-band frequency f_1. When this happens the electric field lines due to the electric coupling and those due to the magnetic coupling of the input directional coupler are as shown in Fig. 36(b). They are orthogonal and 90° out of phase which means that the TE_{111} mode rotates and power may be extracted in the appropriate direction by another directional coupler. At frequencies f_2, remote from the resonant frequency, there is no coupling between directional couplers and hence f_2 input power is transmitted directly to the aerial. Through-losses are as illustrated in Fig. 36(b). It is possible for unwanted modes to upset the cross-loss between f_1 and f_2 but the level of these can generally be reduced by inserting radial or axial rods into the cavity.

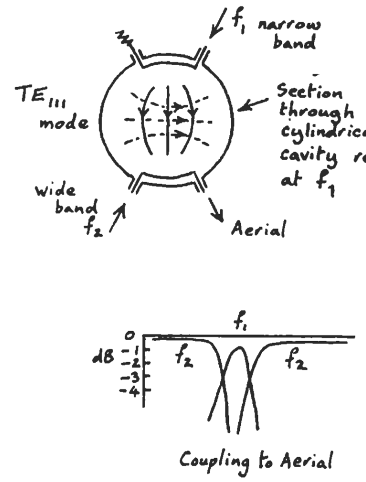
Different f_1 bandwidths may be obtained by using multi-cavity rotamode filters. Figure 37 shows a double-cavity filter where a circular aperture is used as the coupling element.

6.14 Waveguide Combiner for S.H.F.

The circulator, described in Section 5.2, is useful in low-power circuits. One use, where it comes into its own, is for s.h.f. point-to-point transmissions where more than one transmitting frequency is used. A suitable waveguide



(a)



——→—— Electric field due to electric coupling at f_1

– –→– – Electric field due to magnetic coupling at f_1

(b)

Fig. 36. (a) 'Rotamode' combiner arranged as a u.h.f. sound/vision combiner. *(Marconi Communication Systems.)*.

circuit is shown in Fig. 38.

Frequency f_2 is passed by the filter and is transferred by the circulator to the aerial. Frequency f_1 is transferred by the circulator to the filter where it is reflected and transferred to the aerial.

It is necessary for the aerial to be well matched if the cross-loss at frequency f_2 is to remain high, although this can of course be improved by the use of a further filter.

Fig. 37. Double-cavity 'Rotamode' combiner.

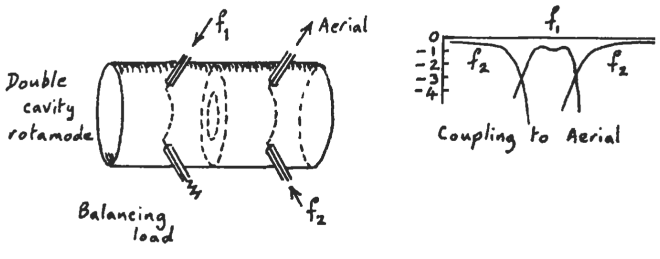
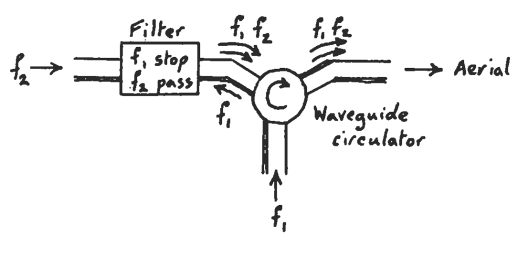


Fig. 38. Waveguide channel combiner.



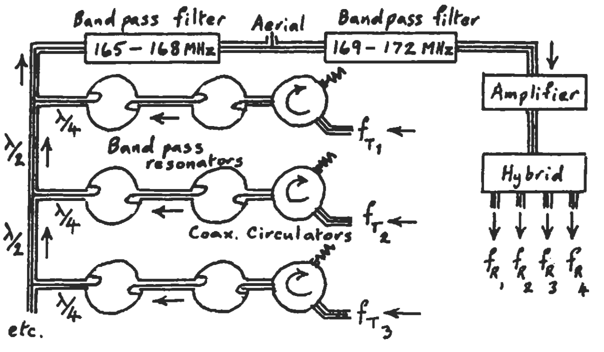


Fig. 39. Combiner using resonators and circulators.

6.15 Combiner for Communication Channels using Resonators and Circulators

A v.h.f. or u.h.f. communications base station often uses a single aerial for both receiving and transmitting several transmissions in the same frequency band. The combined transmitting and receiving aerial system is not used in broadcasting, mainly because different radiation patterns are required for transmitting and receiving. Figure 39 shows the layout of a typical v.h.f. multi-coupled system.

The receiving and transmitting sides are isolated by appropriate filters. On the transmitting side the combiner is effectively a parallel combiner by virtue of the half-wave transmission lines. Band-pass resonators are used in each transmitter chain; circulators are used where there is a problem with low cross-loss and consequent intermodulation products. Any unwanted signals travelling towards a transmitter are directed by the circulator to the circulator load. Wanted signals from the transmitters are transferred to the band-pass filter chain, whence reflections are transferred to the load. Similarly spurious signals from a transmitter are reflected by the band-pass filters into the circulator load.

7 Acknowledgments

The author is indebted to Messrs G. C. Platts and R. M. Ross for making many helpful suggestions during the preparation of this paper and to the Director of Engineering of the British Broadcasting Corporation for giving his permission to publish it.

8 References

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9 Appendix: Early V.H.F. Combiners used by the BBC

9.1 Non-frequency-discriminating Combiner

The first v.h.f. combiner used in the BBC combined the vision and sound frequencies for v.h.f. television. Unlike all the combiners described in Section 6 this combiner contained no resonators or frequency-dependent group of components. As such, it was suitable only for use with an aerial system which had elements fed in phase rotation. The phase progression for one frequency had to be clockwise round the mast and that for the other frequency had to be anti-clockwise. This is sometimes known as space diplexing.

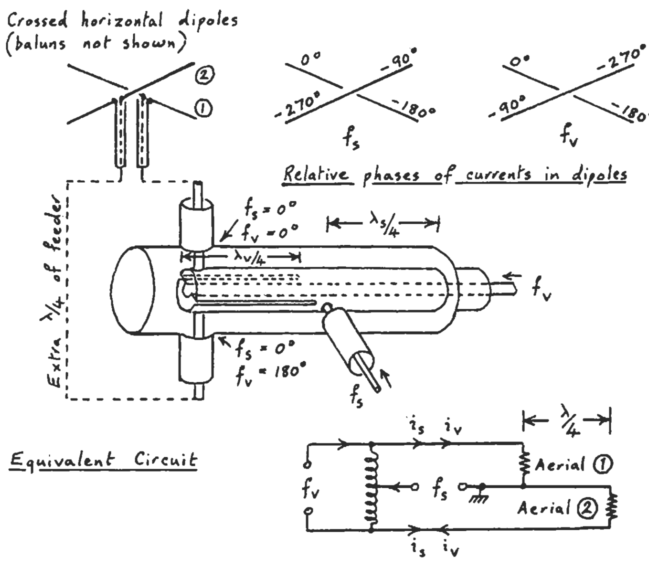

Fig. 40. Non-frequency-discriminating combiner.

The principle of operation and means of achieving the combination in practice are shown in Fig. 40. It will be seen that the combiner is virtually the same as a split-drum hybrid as described in Section 5.1. All that is required to give the necessary phase progression is an extra $\frac{1}{4}$-wavelength on one output feeder. The cross-loss between transmitters, however, is highly dependent on the impedance match of the aerial.

9.2 Coaxial Combiner Directly Derived from Maxwell Bridge

Some of the first combiners for v.h.f. radio channels and v.h.f. sound/vision were of the type illustrated in Fig. 41. It was known that constant impedance combiners were required and it was well known that a lumped circuit Maxwell bridge would perform this function at lower frequencies. The Maxwell bridge, however, has two major disadvantages for high-frequency use. First, it is not possible to earth one side of all components and secondly,

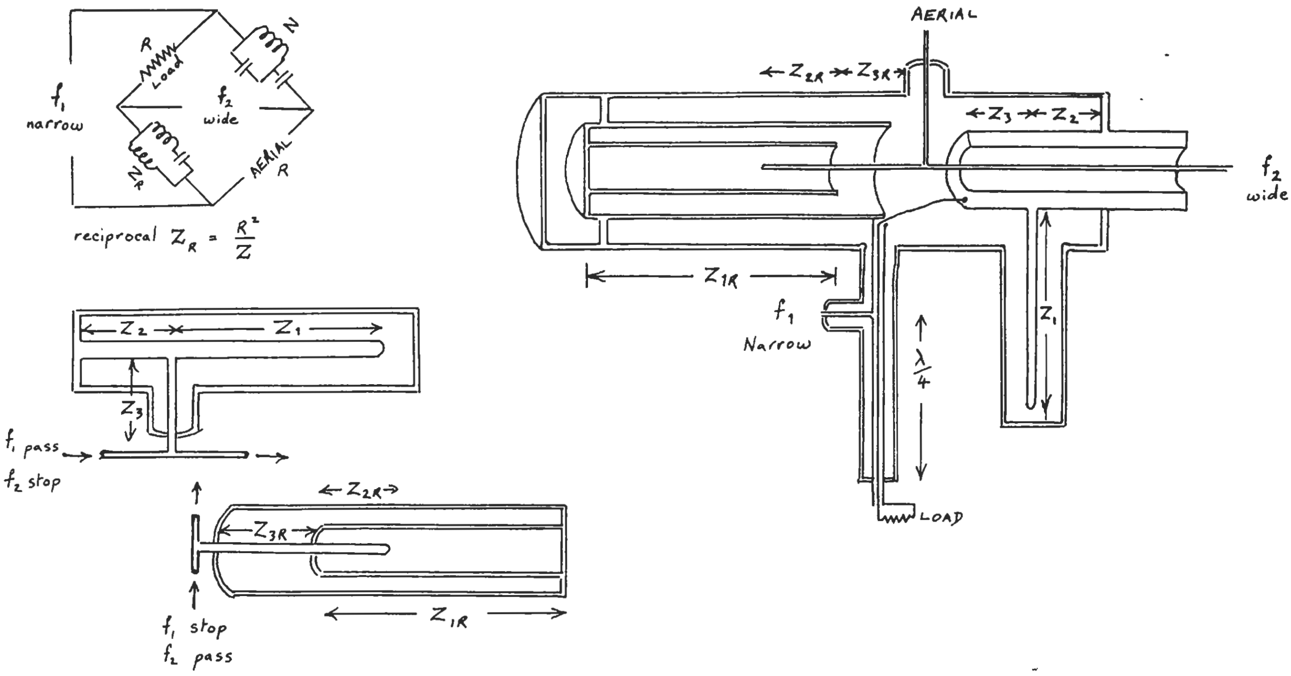

Fig. 41. Maxwell-bridge coaxial combiner.

the two frequency-dependent elements differ and have reciprocal reactances. This means that initial adjustments are complicated.

Mr B. M. Sosin of Marconi was responsible for designing a most ingenious coaxial solution to the problem by using resonators within resonators. He used a coaxial resonator with a transmission line 'stalk' as shown in Fig. 7 on one side of the bridge and then, on the other side, he used series elements instead of parallel elements and vice-versa to achieve the reciprocal reactance. A detailed explanation of the combiner's construction and operation is given in Ref. 6.

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