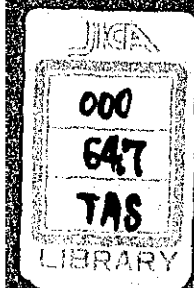


# VHF CIRCUIT THEORY



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## "VHF Circuit Theory"

### 1. Voltage and current in high frequency transmission lines

#### 1.1 When a source end condition is given

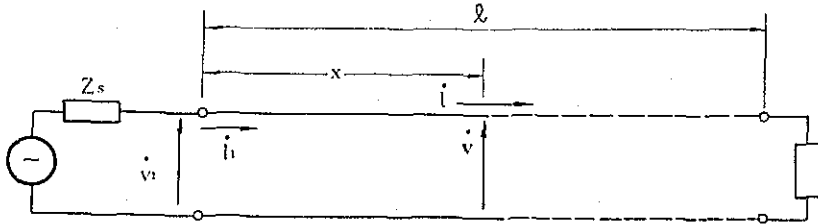


Figure 1.

Voltage  $V$  and current  $I$  at a point  $X$  distance from the source end will be as follows.

$$\dot{V} = \frac{1}{2}(\dot{V}_1 + \dot{I}_1 \dot{Z}_0) e^{-\dot{\gamma}x} + \frac{1}{2}(\dot{V}_1 - \dot{I}_1 \dot{Z}_0) e^{\dot{\gamma}x} \dots\dots\dots (1)$$

$$\dot{I} = \frac{1}{\dot{Z}_0} \left\{ \frac{1}{2}(\dot{V}_1 + \dot{I}_1 \dot{Z}_0) e^{-\dot{\gamma}x} - \frac{1}{2}(\dot{V}_1 - \dot{I}_1 \dot{Z}_0) e^{\dot{\gamma}x} \right\}$$

The first term expresses the incident wave at point  $X$  and the second term the reflected wave.

$\dot{\gamma}$  in the above equation is called the propagation constant.

$$\dot{\gamma} = \sqrt{(R + j\omega L)(G + j\omega C)} = \alpha + j\beta \dots\dots\dots (2)$$

The real part  $\alpha$  of  $\dot{\gamma}$  is called the attenuation constant and its imaginary part  $\beta$  the phase constant.

$R, L, G, C$  here are the resistance, inductance, conductance, and capacitance respectively per unit length of the line.

Also,  $\dot{Z}_0$  in the above equation is called characteristic impedance.

$$\dot{Z}_0 = \sqrt{\frac{(R + j\omega L)}{(G + j\omega C)}} \quad \dots\dots\dots (3)$$

When  $R \ll \omega L$  and  $G \ll \omega C$

$$\begin{cases} \alpha \approx 0 \\ \beta \approx \omega \sqrt{LC} \\ Z_0 = \sqrt{\frac{L}{C}} = \frac{\beta}{\omega C} \end{cases} \quad \dots\dots\dots (4)$$

If the wave length of the line is denoted by  $\lambda$ , the relative wave length by  $k$  and the light velocity by  $c$ , it will be as follows.

$$Z_0 = \frac{\beta}{\omega C} = \frac{2\pi/\lambda}{2\pi f C} = \frac{1}{\lambda f C} = \frac{1}{kcC} \quad \dots\dots\dots (5)$$

Therefore the value of  $Z_0$  may be determined if the capacitance per unit length and the relative wave length are known.

Equation (1) denotes the incident wave of  $|1/2 (\dot{V}_1 + \dot{I}_1 \dot{Z}_0)|$  at the source end and the reflected wave of  $|1/2 (\dot{V}_1 - \dot{I}_1 \dot{Z}_0)|$  also at the source end, and shows that the amplitude of the incident wave decreases by  $e^{-\alpha x}$  times at  $x$  distance from the source end and that the reflected wave increases by  $e^{\alpha x}$  times. Although the foregoing is in relation to the voltage, as current may also be determined in the same manner, the same may be said of both the incident wave and reflected wave.

#### 1.2 When a load end condition is given

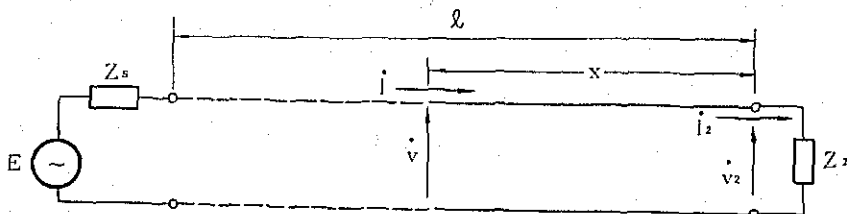


Figure 2.

Voltage  $V$  and current  $I$  at a point  $x$  distance from the load end will be as follows.

$$\dot{V} = \frac{1}{2}(\dot{V}_2 + \dot{I}_2 Z_0) e^{\dot{\gamma}x} + \frac{1}{2}(\dot{V}_2 - \dot{I}_2 Z_0) e^{-\dot{\gamma}x} \dots\dots\dots (6)$$

$$\dot{I} = \frac{1}{Z_0} \left\{ \frac{1}{2}(\dot{V}_2 + \dot{I}_2 Z_0) e^{\dot{\gamma}x} - \frac{1}{2}(\dot{V}_2 - \dot{I}_2 Z_0) e^{-\dot{\gamma}x} \right\}$$

The first term expresses the incident wave at point  $x$  and the second term the reflected wave. Although the above are of the same form as equation (1), the direction of  $x$  and the sign of  $\dot{\gamma}$  are different.

The ratio  $\dot{\Gamma}$  of the reflected wave to the incident wave is called the reflection coefficient.

$$\dot{\Gamma} = \frac{V_2 - I_2 Z_0}{V_2 + I_2 Z_0} = \frac{V_2/I_2 - Z_0}{V_2/I_2 + Z_0} = \frac{Z_2 - Z_0}{Z_2 + Z_0} \dots\dots\dots (7)$$

In relation to power, when a power of 1 progresses from the source,  $|\dot{\Gamma}|^2$  is reflected and a power of  $1 - |\dot{\Gamma}|^2$  is absorbed in the load.

The ratio  $\rho$  of the voltage at the point where the incident wave and the reflected wave overlaps to the voltage at the point where the incident wave and the reflected wave cancel each other out is called the standing wave ratio.

$$\rho = \frac{|V_2 + I_2 Z_0| + |V_2 - I_2 Z_0|}{|V_2 + I_2 Z_0| - |V_2 - I_2 Z_0|} = \frac{1 + |\dot{\Gamma}|}{1 - |\dot{\Gamma}|} \dots\dots\dots (8)$$

When  $Z_2$  is the pure resistance  $R_2$

$$\rho = \frac{1 + \frac{R_2 - Z_0}{R_2 + Z_0}}{1 - \frac{R_2 - Z_0}{R_2 + Z_0}} = \frac{R_2}{Z_0} \dots\dots\dots (9)$$

### 1.3 Input Impedance

To determine how input impedance  $Z_{in}$  will appear when load  $Z_2$  is applied to a line of length  $x$ , since  $V$  and  $I$  have now

been determined, the ratio of these two factors will be  $Z_{in}$ . In other words, it will be as follows.

$$\dot{Z}_{in} = \dot{Z}_0 \frac{\frac{\dot{Z}_2}{\dot{Z}_0} + \tanh \dot{\gamma}x}{1 + \frac{\dot{Z}_2}{\dot{Z}_0} \tanh \dot{\gamma}x} \dots\dots\dots (10)$$

If this line is lossless, the above equation will be as follows as  $\alpha = 0$  in the equation  $\dot{\gamma} = \alpha + j\beta$

$$\dot{Z}_{in} = \dot{Z}_0 \frac{\frac{\dot{Z}_2}{\dot{Z}_0} + j \tan \beta x}{1 + j \frac{\dot{Z}_2}{\dot{Z}_0} \tan \beta x} \dots\dots\dots (11)$$

We shall next use an example to explain what happens to  $Z_{in}$  when  $Z_2$  and  $\beta x$  take a certain value.

- (1) When  $Z_2 = 0$ , it will be as follows.

$$Z_{ins} = -jZ_0 \tan \beta x \dots\dots\dots (12)$$

In other words, if the load end is short circuited so it is shorter than  $\lambda/4$ , it will become inductive.

Also, when  $Z_2 = \infty$ , it will be

$$Z_{ino} = -jZ_0 \cot \beta x \dots\dots\dots (13)$$

Therefore, regardless of the length of the line, it will be

$$Z_{ins} \cdot Z_{ino} = Z_0^2 \dots\dots\dots (14)$$

- (2)  $Z_{in}$  at  $\beta x = \beta x_1$  and  $Z_{in}$  at  $\beta x = \beta (x_1 + 0.5\lambda)$  are the same. That is, if  $x$  is increased, same value of  $Z_{in}$  will appear at every  $0.5\lambda$ .



(3) When  $\beta x = \frac{\pi}{2}$  ( $x = 0.25\lambda$ ), it will be as follows

$$Z_{in} = \frac{Z_0^2}{Z_2} \dots\dots\dots (15)$$

In other words, the larger the value of  $Z_2$  the smaller the input impedance  $Z_{in}$  at  $\lambda/4$  line.

The above can also be confirmed by means of the Smith Chart which will be explained later.

1.4 The composition of the Smith Chart and method of usage of it  
Calculations to obtain input impedance  $Z_{in}$  by means of equation (11) and reflection coefficient by means of equation (7) are very complicated. The Smith Chart was prepared to simplify this calculation.

(Examples of usage)

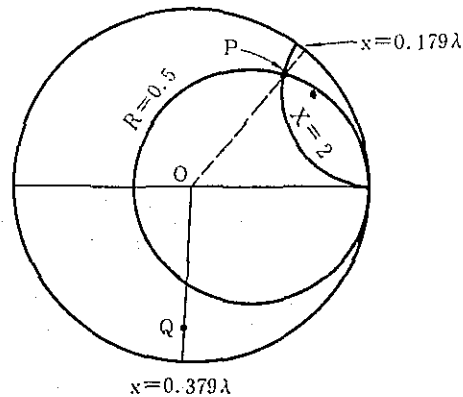
Load impedance

$$(25 + j100) \text{ ohm}$$

Characteristics impedance

$$50 \text{ ohm}$$

Obtain source end input  
impedance with line  
length at  $3.2 \lambda$ .



(1) First, obtain normalized impedance  $Z_n$ .

$$Z_n = \frac{1}{50}(25 + j100) = 0.5 + j2 \dots\dots\dots (15)$$

(2) Obtain reflection coefficient  $\Gamma_2$  of the load end when load is  $0.5 + j2$ .

Find circles with  $R = 0.5$  and  $X = +2$  from the chart and denote their point of intersection as P. Then, OP will be  $\Gamma_2$ .

- (3) Obtain phase angle of  $\Gamma_2$ .

Extend OP and read distance scale at the point where it intersects the large circle to obtain  $0.179\lambda$ .

Note: This 0.179 is not the amount of phase  $\Gamma_2$  indicated directly but is the practical value given to enable readings to be taken with the position of  $\Gamma_2$  as the standard as  $\Gamma_x$  being obtained next will determine the position by advancing the dial by a certain number of wave lengths.

The true phase angle of  $\Gamma_2$  is, in fact, accurately indicated by the actual angle drawn. As angle scales are indicated in combination with distance scales, if readings are taken,  $\Gamma_2$  will be  $51^\circ$ .

If you will recall from equation (7) that

$$\Gamma_2 = \frac{Z_n - 1}{Z_n + 1}$$

and calculations were made in this case, it will be

$$\begin{aligned}\Gamma_2 &= \frac{0.5 + j2 - 1}{0.5 + j2 + 1} = \frac{-0.5 + j2}{1.5 + j2} = \frac{3.25}{6.25} + j\frac{4}{6.25} \\ &= 0.52 + j0.64 \quad \dots\dots\dots (16)\end{aligned}$$

$$\text{Phase angle of } \Gamma_2 = \tan^{-1}(0.64/0.52) = 51^\circ$$

thus proving the above reading.

- (4) Obtain  $\Gamma_x$

As  $x$  is  $3.2\lambda$ , we shall consider it as  $3\lambda + 0.2$  but it will not be necessary to consider the  $3\lambda$  portion as this portion will simply be the vector making 6 revolutions and returning to its original position. Consider the  $0.2\lambda$  portion only and, adding this  $0.2\lambda$  to the  $0.179\lambda$  of  $\Gamma_2$ , turn clock-wise to locate and determine the position of  $0.379\lambda$ . If we draw line from this point to

the center of the Chart, the angle formed will be the angle of  $\Gamma_x$ . If we now set point Q on the straight line so that  $OQ = OP$ , then  $OQ$  will be  $\Gamma_x$ .

- (5) Obtaining normalized input impedance  $Z_{nx}$ .

It will only be necessary to determine the parameter of the real circle passing through point Q and the parameter of the imaginary circle. Considering these to be in proportion in this instance, we shall read Circle R as 0.175 and Circle X as a capacitive 0.94.

$$Z_{nx} = 0.175 - j0.94 \quad \dots\dots\dots (17)$$

- (6) Convert to input impedance  $Z_{in}$ .

Simply multiply by the characteristics impedance of 50 ohm.

$$Z_{in} = 8.75 - j47 \quad \dots\dots\dots (18)$$

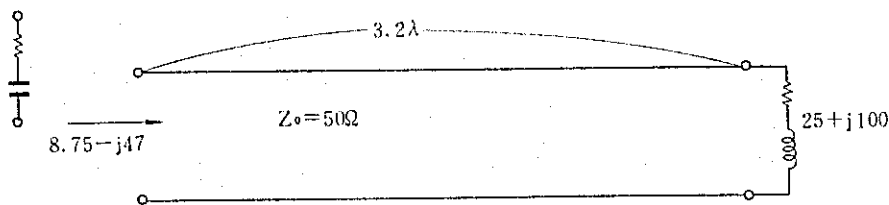


Figure 4.

### 1.5 Tendency of input impedance to approach $Z_0$ due to loss.

In lines with losses,  $\Gamma_x$  will decrease at the rate of  $e^{-2\alpha x}$ . In large  $\alpha$  and  $x$  transmission lines, input impedance will appear as  $Z_0$  regardless of the load applied.

With reference to  $\Gamma_x$  in the Smith Chart, as the position of  $\Gamma^2$  will rotate clock-wise one revolution each time  $x$  advances  $0.5\lambda$ , the vector will have to be rotated scores

of times to obtain  $\Gamma_x$  if  $x$  happens to be several hundred meters.

As the important thing is that  $\Gamma_x$  decreases at the rate of  $e^{-2\alpha x}$  as  $x$  advances,  $\Gamma_x$  will steadily approach its origin of (1,0) and, if  $\alpha x$  is large, it will finally converge at its point of origin. The input impedance at this time will be  $Z_0$  itself.

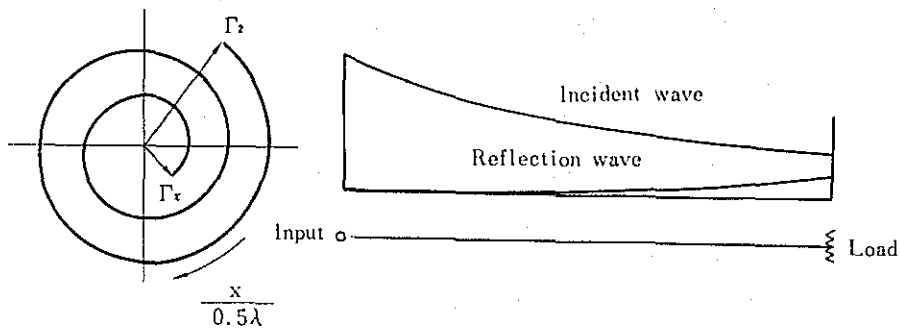


Fig. 5.

#### 1.6 The possibility of ghost images being generated due to mis-match.

If a voltage is impressed on a transmission line, it will travel in the direction of the load in the form of a incident wave. If the load is mis-matched, a portion of this will be reflected back to the source. The time required will be the length of path over and back divided by the propagation velocity. The source condition will be subject to change the instant the reflection wave returns to the input. Moreover, if the source impedance is mis-matched, a portion of the reflection wave that has returned will be reflected back again as a part of the incident wave and be reflected again at the load end. This action will continue to repeat itself.

We shall now use equation (6) and express the voltage at

the source end as a large cluster of reflection waves.  
According to equation (6) impedance  $Z_{in}$  as seen from the  
source end will be as follows.

$$Z_{in} = \frac{\dot{V}_1}{\dot{I}_1} = Z_0 \frac{(V_2 + I_2 Z_0)e^{\gamma l} + (V_2 - I_2 Z_0)e^{-\gamma l}}{(V_2 + I_2 Z_0)e^{\gamma l} - (V_2 - I_2 Z_0)e^{-\gamma l}}$$

$$= Z_0 \frac{(Z_2 + Z_0) + (Z_2 - Z_0)e^{-2\gamma l}}{(Z_2 + Z_0) - (Z_2 - Z_0)e^{-2\gamma l}} \dots\dots\dots (19)$$

Voltage  $V_1$  at the source end will be

$$V_1 = E \frac{Z_{in}}{Z_s + Z_{in}} = E \frac{Z_0 \frac{(Z_2 + Z_0) + (Z_2 - Z_0)e^{-2\gamma l}}{(Z_2 + Z_0) - (Z_2 - Z_0)e^{-2\gamma l}}}{Z_s + Z_0 \frac{(Z_2 + Z_0) + (Z_2 - Z_0)e^{-2\gamma l}}{(Z_2 + Z_0) - (Z_2 - Z_0)e^{-2\gamma l}}}$$

$$= E \frac{Z_0(Z_2 + Z_0) + Z_0(Z_2 - Z_0)e^{-2\gamma l}}{(Z_s + Z_0)(Z_2 + Z_0) - (Z_s - Z_0)(Z_2 - Z_0)e^{-2\gamma l}}$$

$$= E \frac{Z_0}{Z_s + Z_0} \left\{ 1 + \frac{Z_2 - Z_0}{Z_2 + Z_0} e^{-2\gamma l} \right\} \times \frac{1}{1 - \frac{(Z_s - Z_0)(Z_2 - Z_0)}{(Z_s + Z_0)(Z_2 + Z_0)} e^{-2\gamma l}}$$

$$= E \frac{Z_0}{Z_s + Z_0} \left\{ 1 + \frac{Z_2 - Z_0}{Z_2 + Z_0} e^{-2\gamma l} \right\} \sum_{n=0}^{\infty} \left\{ \frac{(Z_s - Z_0)(Z_2 - Z_0)}{(Z_s + Z_0)(Z_2 + Z_0)} e^{-2\gamma l} \right\}^n$$

\dots\dots\dots (20)

In equation (20),  $E \frac{Z_0}{Z_s + Z_0}$  indicates the first incident

wave,  $\frac{Z_2 - Z_0}{Z_2 + Z_0} e^{-2\gamma \ell}$  indicates the reflected wave of the first incident wave that has reached the load end,  $\frac{(Z_s - Z_0)(Z_2 - Z_0)}{(Z_s + Z_0)(Z_2 + Z_0)} e^{-2\gamma \ell}$  indicates waves that have been reflected back from both the load end and the source end, and  $\sum_{n=0}^{\infty}$  indicates the number of times this has been repeated.

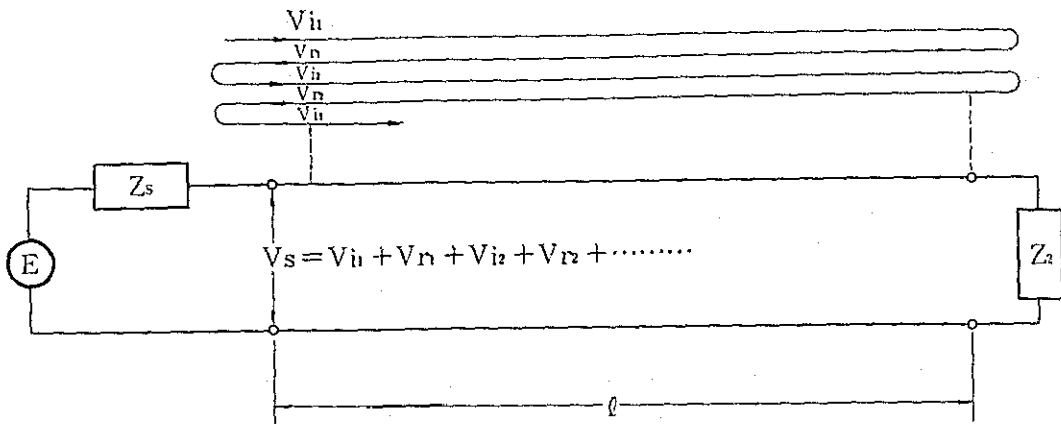


Fig. 6

## 2. Practical example of a distributed-constant circuit

### 2.1 Feeder

Unbalanced coaxial type copper tubes and coaxial cables are available for use as feeders for TV and FM broadcasts. Waveguides are used particularly in the case of high power broadcasts.

#### (1) Characteristic impedance

Characteristic impedance ( $Z_0$ ) of coaxial feeders may be expressed by the following equation.

$$Z_0 = \frac{138}{\sqrt{\epsilon}} \log_{10} \frac{D}{d} (\Omega) \quad \dots\dots\dots (21)$$

Here,  $\epsilon$  is the average dielectric constant of the insulation,  $D$  the internal diameter of the outer conductor and  $d$  the outer diameter of the inner conductor.

(2) Attenuation

Resistance attenuation  $\alpha_r$  may be expressed by the following equation.

$$\alpha_r = \frac{0.0629 \sqrt{\epsilon} \sqrt{f} \frac{\sqrt{\rho_1}}{d} + \frac{\sqrt{\rho_2}}{D}}{\log_{10} \frac{D}{d}} \text{ (dB/km)} \dots\dots\dots (22)$$

$\rho_1$  and  $\rho_2$  are the resistivity of the outer and inner conductors. The attenuation per 100 m of coaxial cables in general use at a temperature of  $20^\circ$  is shown in Figure 7.

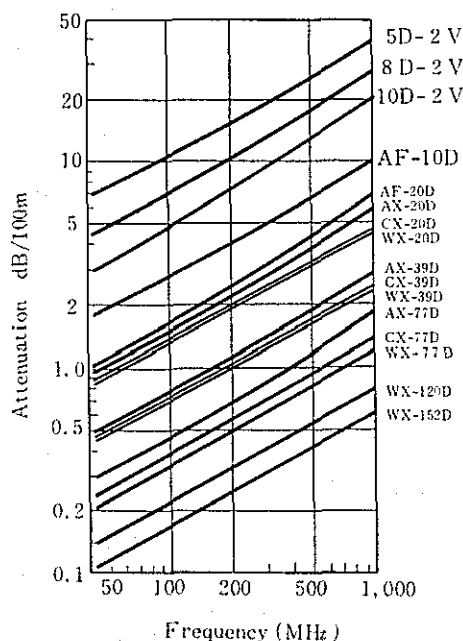


Figure 7. Standard Attenuation of Feeders

### (3) Allowable Power

The maximum allowable power of coaxial feeders is determined by the deformation of the internal conductor insulation by heat rise due to power loss above the VHF band. Considering the safety factor, the internal conductor temperature rise in feeders for broadcast use is held to 25°C when the external temperature is 50°C. The allowable power for coaxial feeders in general use is shown in Figure 8.

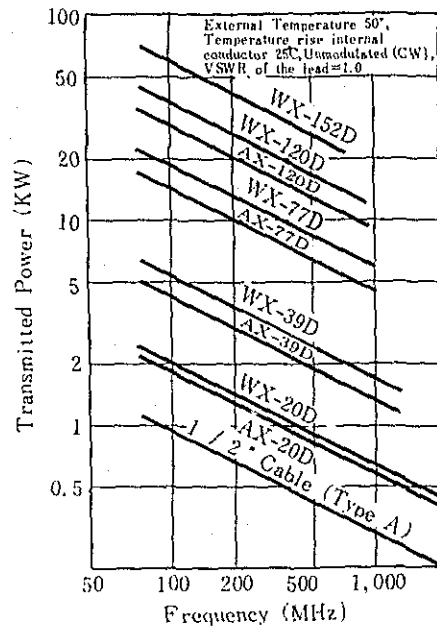


Figure 8. Allowable Power of Feeders

Table 1. Dimension and Characteristics of Coaxial Copper Tube Feeders

Type	Outer Conductor		Inner Conductor		Standard Length (m)	Characteristics		
	External Diameter (m)	Internal Diameter (m)	External Diameter (m)	Internal Diameter (m)		Impedance	TE "mode cut-off frequency (GHz)	Standard Value of DC Resistance (Ω/100m)
WX-152D	155.6	151.9	66.0	64.0	6,000	50	0.9	0.011
WX-120D	123.2	120.0	52.1	50.1		50	1.13	0.014
WX- 77D	79.4	76.9	33.4	31.3		50	1.77	0.022
WX- 39D	41.3	38.8	16.9	14.9	6,122	50	3.50	0.045
WX- 20D	22.22	19.94	8.66	7.39	3,000	50	6.82	0.13



## 2.2 Multiplex Feed

There are various methods of multiplex feed depending on conditions such as (1) bandwidth of the wave to be fed, (2) relations with the frequency of other waves and (3) the power used. Duplex feed systems are comprised principally of bridge diplexers (BD), constant impedance notch diplexers (CIN) and notch diplexers (ND). Multiplex feed systems are composed of multiple use of these systems. Bridge diplexers may be used regardless of the bandwidth of each wave and the frequency separation between two waves as long as it is within its usable band. The constant impedance notch diplexer is suitable for use where the frequency separation is comparatively narrow while the notch diplexer is suitable where the frequency separation is comparatively wide.

### (1) CIN

In Figure 9, input  $f_1$  is divided into two parts in hybrid circuit #1, passes through the filters ( $f_1$  pass,  $f_2$  stop type) with no loss in frequency response, and is combined in hybrid circuit #2 and is transmitted to the antenna. Input  $f_2$  is divided into two parts in hybrid circuit #2, reflected in the filters with no loss in its and is transmitted to the antenna after synthesizing again in hybrid circuit #2. Therefore, the transmission characteristic of input  $f_1$  to the antenna is practically equal to the pass characteristic of the filter, and the transmission characteristic of input  $f_2$  is practically equal to the reflection characteristic of the filter. Also, in relation to leakage between the various inputs, although leakage from input  $f_2$  to input  $f_1$  will pose no problems, leakage from input  $f_1$  to input  $f_2$  will be determined by the characteristics of the #2 hybrid circuit.

Hybrid circuits currently in wide use are bridge diplexer,

ratrace, hybrid ring  
(2 element or multi-  
element type), 3dB  
coupler and shortslot  
circuits.

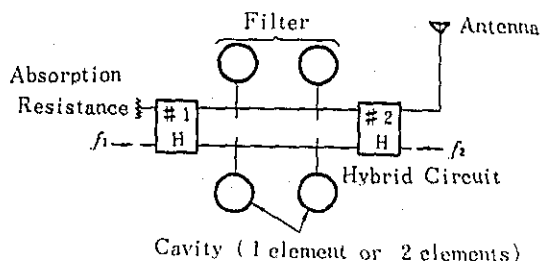


Figure 9. CIN Type Duplex  
Feed System

(2) Video and Sound Multiplexing CIN

The video pass characteristic  $\alpha_T$  of 1 element type  
filters employed in video and sound multiplexing circuits  
as shown in Figure 10 may be expressed as in equation (23)  
and in Figure 10.

$$\alpha_T = 10 \log_{10} \left( 1 + \frac{1}{4} y^2 \right) \text{dB} \quad \dots\dots\dots (23)$$

$$y = a \frac{f - f_v}{f - f_a}$$

$a$  = Constant

$f_v$  = Video carrier frequency

$f_a$  = Sound carrier frequency

Due to video envelope delay characteristic and sound  
reflection loss here, at a video carrier frequency  
+4.2MHz -1.5dB is considered standard as the video pass  
characteristics. Constant in equation (23) is determined  
from the foregoing.

A cavity resonant with sound is used as the filter, and  
when this cavity is connected to the line (loaded)  $Q$  of  
this circuit, loaded  $Q$ ,  $Q_L$  may be expressed by equation  
(24).

$$Q_L = \frac{f_a}{|a|(f_a - f_v)} \dots\dots\dots (24)$$

Although the attenuation of the video carrier frequency in equation (23) will become 0dB, this is due to the fact that the circuit is a pure reactance circuit. If we consider resistance losses, loss  $\alpha_{Tfv}$  of the video carrier frequency may be expressed by equation (25).

$$\alpha_{Tfv} = 10 \log_{10} \left( 1 + \frac{1}{2\delta^2 Q_O Q_L} \right) \text{dB} \dots\dots\dots (25)$$

$Q_O$  = Unloaded Q of cavity

$$\delta = \frac{f_a - f_v}{f_a}$$

Also, reflection characteristic  $\alpha_r$  of sound may be expressed by equation (26).

$$\alpha_r = 10 \log_{10} \left( 1 + \frac{4}{y^2} \right) \text{dB} \dots\dots\dots (26)$$

If we consider resistance losses in the sound carrier frequency, it may be expressed by equation (27)

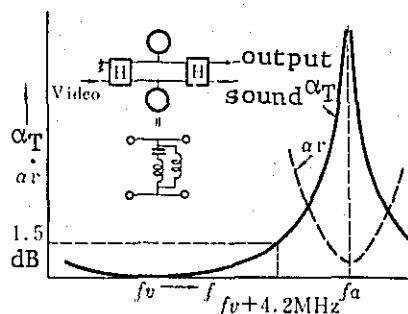


Figure 10. CIN Pass and Reflection Characteristics

$$\alpha_{rfa} = 20 \log_{10} \left( 1 + \frac{Q_L}{Q_0} \right) \text{dB} \quad \dots\dots\dots (27)$$

The losses expressed in equations (25) and (27) become heat losses and causes temperature rise in the cavity. As  $\alpha_{Tfv}$  is small in relation to  $\alpha_{rfa}$ , cavity heat is believed due to reflection loss of the sound power. As the cavity must be thoroughly stable against to this heat, a cavity will be required to hold temperature rise to approximately 25°C in relation to the external temperature and to be in accordance with sound power.

### (3) Notch Diplexer

Figure 11 is an example of a notch diplexer layout.

Filter  $F_1$  is of the  $f_1$  pass and  $f_2$  block type and filter  $F_2$  is of the  $f_2$  pass and  $f_1$  block type.

Input  $f_1$  passes through filter  $F_1$  without loss in its band and is reflected (blocked) by the filter  $F_2$  and is transmitted to

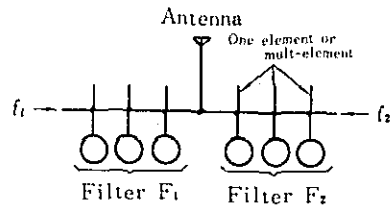


Figure 11. Notch Diplexer Type Duplex Feed System

the antenna. Similarly, input  $f_2$  is reflected by filter  $F_1$  after passing through filter  $F_2$  and is transmitted to the antenna. Filters  $F_1$  and  $F_2$  will be of one element or multi-element composition according to the frequency separation of  $f_1$  and  $f_2$  and the bandwidth.

## 2.3 Resonator

### (1) Double Coaxial Type Resonator

The theoretical construction of a resonator will be as

shown in Figure 12. C for parallel resonance is sometimes in the form of lumped constants or may use the capacitance between the wall but, in any event, it is arranged so that fine adjustments are possible from the outside. If the electrical length of the respective characteristic impedances  $W_1$  and  $W_2$  of the outer and inner coaxial tubes is designated as  $\theta_{1s}$  and  $\theta_{2s}$  at series resonant frequency  $f_s$ , it will be as follows.

$$W_1 \cdot \tan \theta_{1s} + W_2 \tan \theta_{2s} = 0 \quad \dots\dots\dots (28)$$

Also, at the parallel resonant frequency  $f_o$  it will be

$$W_1 \cdot \tan\left(\theta_{1s} \frac{f_o}{f_s}\right) + W_2 \tan\left(\theta_{2s} \frac{f_o}{f_s}\right) = X_c \quad \dots\dots\dots (29)$$

As the  $f_s$  of each element and the size of parallel C will be determined by the method in which the series parallel resonant frequency is applied, the length of the coaxial tube of each of the element may be designed by using these values in the above equation. Further, if the shift of  $\theta_{1s}$  and  $\theta_{2s}$  from  $\frac{\pi}{2}$  in this calculation is designated as  $\Delta\theta_{1s}$  and  $\Delta\theta_{2s}$ , it will be sufficient to use approximate equations such as  $\tan \theta_{1s} \approx -\frac{1}{\Delta\theta_{1s}}$  and  $\tan \theta_{2s} \approx \frac{1}{\Delta\theta_{2s}}$ .

Although there are only two free degrees left of the 4 unknown elements in the above two equations, the value of W is selected at approximately 35 to 45Ω and, as short circuit current normally rises sharply if  $\theta$  is near 90 degrees, it will be satisfactory to select 80 degrees and 100 degrees as the value of  $\theta$ . As there is very little energy other than the carrier wave, it will only be necessary to consider the carrier wave when calculating KVA. If we use Figure 13 as a reference and consider the example a 1kW unit, it will be as follows as it may be divided into two equal parts by means of a bridge.

$$P = \frac{1000}{2} = 500W \quad E = \sqrt{500W \times 50\Omega} = 158V \quad \dots\dots (30)$$

If  $X_c = 75\Omega$ , it will be

$$I_c = I_L = \frac{158}{75} = 2.1A \quad \dots\dots\dots (31)$$

If  $\theta = 100$  degrees,

$$I_s = \frac{I_L}{\cos\theta} = \frac{2.1}{0.17} = 12.3A \quad \dots\dots\dots (32)$$

Maximum voltage is generated at 90 degrees from the end

$$V = 50\Omega \times 12.3A = 615V$$

In the case of withstanding voltages, we must consider  $\sqrt{2}$  times of this value.

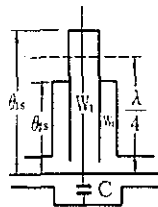


Figure 12. Construction of a resonator

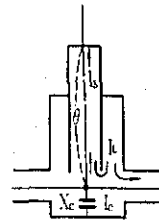


Figure 13. Calculation of the current

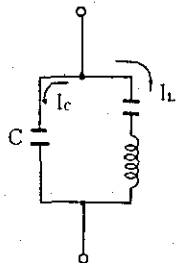


Figure 14. Equivalent Circuit

## (2) High Q Resonator

When we combine the video transmitter output and the sound transmitter output in UHF, a high Q sound reflector will be necessary.

The sound reflector is an open end or shorted end cavity as shown in Figure 15. This type of construction was used to obtain very high Q and in this example the Q is over 10,000. As indicated by the equivalent circuit in Figure 16, the resonant point of this cavity is determined by  $L_2$  and  $C_2$  which are, in turn, determined by the length of the center cylinder of the cavity. Fine adjustments of the frequency may be carried out by means of the piston protruding from the cavity.

Also, as the parallel resonant point is at a lower point than the series resonant point and correspond to video carrier frequencies, parallel resonance is realized by shortening a short circuited end coaxial tube from  $\lambda/4$  and operating it as an equivalent inductance.

As the Q is high in sound resonators, close attention must be paid to any changes in resonant frequency due to temperature. This becomes increasingly important as the frequency becomes higher. For example, if the resonant frequency of the sound resonator shifts due to temperature changes, sound reflections will become uncomplete and the waves will be absorbed in the absorption resistor of the bridge connected to the input side of this unit. In other words, diplexer functions will become imperfect and a part of the sound power will not reach the antenna or, in color system, fluctuations in the sound resonator notch will affect the components in the vicinity of the subcarrier and thus deteriorate the color characteristics.

Due to these reasons, Invar, which has a particularly low

temperature coefficient is used for the center conductor and steel is used for the outer conductor to reduce resonant frequency fluctuations in this reflector.

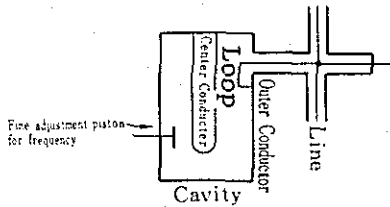


Figure 15. Construction of a Sound Reflector

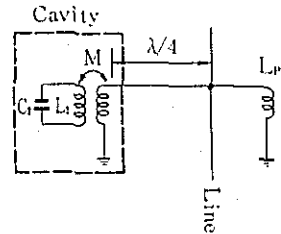


Figure 16. Equivalent Circuit of a Sound Reflector

#### 2.4 Bridge Diplexer

The bridge diplexer is one form of hybrid circuit and its construction is shown in Figure 17(a). As leakage between ports 1 and 3 are determined by the slit width and the dimension of the inner and outer conductors, these leakage may be disregarded as long as mechanical troubles do not develop in the finished product. Its equivalent circuit will be as shown in Figure 17(b). In this figure,  $N_1N_3$  is the network originating from this construction and properties of bridge diplexer may be easily explained by this equivalent circuit. In other words, waves entering from port 1 will not appear at port 3 but will proceed to ports 2 and 4 in antiphase. Also, waves entering from port 3 will not go to port 1 but will proceed to 2 and 4 in in-phase. Waves that enter from ports 2 and 4 with the phase of 180 degrees will not go to port 3 but will proceed to port 1 only and waves that enter in-phase proceed to port 3 only.



If the phase between two waves which enter from ports 2 and 4 exist during those values, the amplitude of the wave at 1, 3 will change in accordance with this phase. Therefore, regardless of the type of load at 2 and 4, as long as they are equal, the reflected waves of the load by the input wave from port 1 will always return to port 1 if the load is equal-distant from 2 and 4, and, if the difference in the distance of the load is  $\lambda/4$ , the reflected wave will proceed to port 3 only. The load reflection wave by the input wave from port 3 will be just the opposite as it will return to 3 when the difference in distance is 0 and will appear at port 1 if the difference is  $\lambda/4$ . Therefore, if the load is shorted out or becomes  $50\Omega$  according to the frequency, the load should be connected to  $\lambda/4$  difference to cause the reflection wave to be absorbed in port 3.

If connections are made in this manner, the input wave from 1 will proceed with no reflections. In other words, this will mean that the circuit will be matched. When the load is  $50\Omega$  as in the case of the antenna, a resistance of  $100\Omega$  will exist in series between 2 and 4 and this will be stepped down by a 1 : 2 ideal transformer to become  $25\Omega$  on the primary side. To match this to the  $50\Omega$  input impedance in the primary side, it will become necessary to insert a transformer.

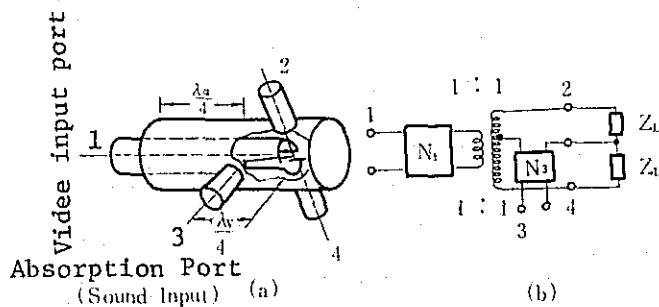


Figure 17. Split Coaxial Type Bridge

## 2.5 Impedance Transformer

Figure 18 shows matching conditions of resistor  $R$  and the feeder of characteristic impedance  $W_0$ .

Figure 19 shows the impedance transformation characteristics of a series type matching circuit.

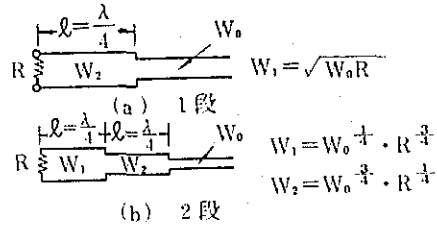


Figure 18. Matching Conditions of a Series Type Matching Circuit

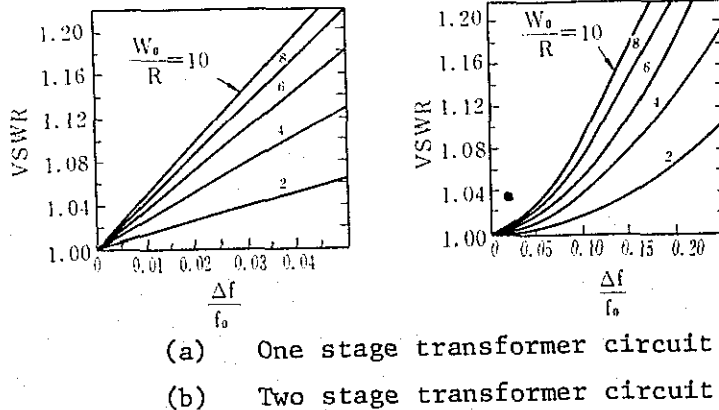


Figure 19. VSWR Frequency Characteristics of a Transformer Circuit

## 2.6 Power Dividing Circuit

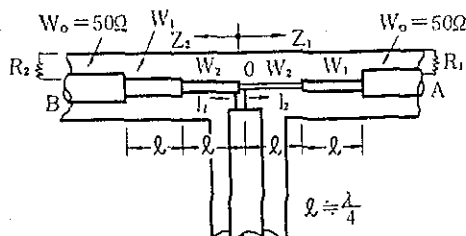
With loads  $R_1$  and  $R_2$  connected to terminals A and B in the power dividing circuit in Figure 20 both at  $50\Omega$ , designate  $Z_1$  as the impedance seen from point "O" in the direction of load  $R_1$  and  $Z_2$  as the impedance seen from point "O" in the direction of load  $R_2$ . Also, if the power dividing ratio supplied to the load is designated as  $m/n$ , the following relations will exist between the power dividing ratio and

$Z_1, Z_2$ .

$$\begin{aligned} Z_1 &= \frac{m+n}{n} W_o = \frac{500}{n} (\Omega) \\ Z_2 &= \frac{m+n}{m} W_o = \frac{500}{m} (\Omega) \end{aligned} \quad ) \dots\dots\dots (33)$$

However, here  $m + n = 10$

The matched circuit of  $R_1$  to  $Z_1$  and  $R_2$  to  $Z_2$  can be obtained from the matching conditions in Figure 18.



(b) Power dividing circuit

Figure 20. T Type Circuit and Power Dividing Circuit

## 2.7 Directional Coupler

Directional couplers are used in high frequency coupling and Figure 21 shows a theoretical diagram of a directional coupler. As is shown in the diagram, a hole is provided in the outer tube of the coaxial feeder through which a rotatable loop is inserted. One end of this loop is terminated with a resistor and high frequency waves are drawn from the other end. In this case, if the voltage and current at a point on the line are designated as  $V$  and  $I$ , and, if this were to be divided into incident waves  $V_o$  and  $I_o$  and reflection waves  $V_o'$  and  $I_o'$ , the following relations will exist.

$$V = V_o + V_o'$$

$$I = I_o - I_o'$$

$$V_o = Z_o I_o, V_o' = Z_o I_o' \quad \dots\dots\dots (34)$$

$$V = Z_o (I_o + I_o')$$

If capacitive coupling is now made to the line, amounts proportional to voltage V can be obtained and, if inductive coupling is made, amounts proportional to current I can be obtained.

Voltage V is divided by capacitor C and resistor R and, if

$$\frac{1}{\omega C} \gg R$$

then, voltage  $e_c$  developed at the ends of resistor R is

$$e_c = j\omega CRV = j\omega CRZ_o (I_o + I_o') \quad \dots\dots\dots (35)$$

Voltage  $e_M$  obtained by M coupling will be

$$e_M = j\omega MI = j\omega M (I_o - I_o') \quad \dots\dots\dots (36)$$

If coefficients  $j\omega CRZ_o$  and  $j\omega M$  in the above two equations were made equal, it will be

$$j\omega CRZ_o = j\omega M = K$$

$$e_c = K(I_o + I_o') \quad \dots\dots\dots (37)$$

$$e_M = K(I_o - I_o')$$

If we take the sum and difference of these two voltages, it will be

$$e_c + e_M = 2KI_o \quad \dots\dots\dots (38)$$

$$e_c - e_M = 2KI_o'$$

As the sum voltages will indicate incident waves and difference voltages the reflection waves, if this were detected

and indicated, it may be used as a incident - reflection wattmeter.

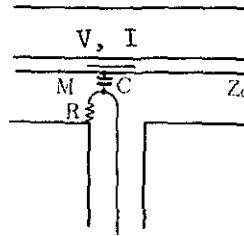


Figure 21. CM Type Directional Coupler

## 2.8 Balun

When connecting unbalanced circuits such as coaxial feeders and balanced circuits such as parallel feeders, it will be necessary to employ a balun. There are various types of baluns such as (a) split balun, (b) U-balun, (c) Sperrtop and (d) stub. Baluns serve to block the current in the outer part of the coaxial feeder.

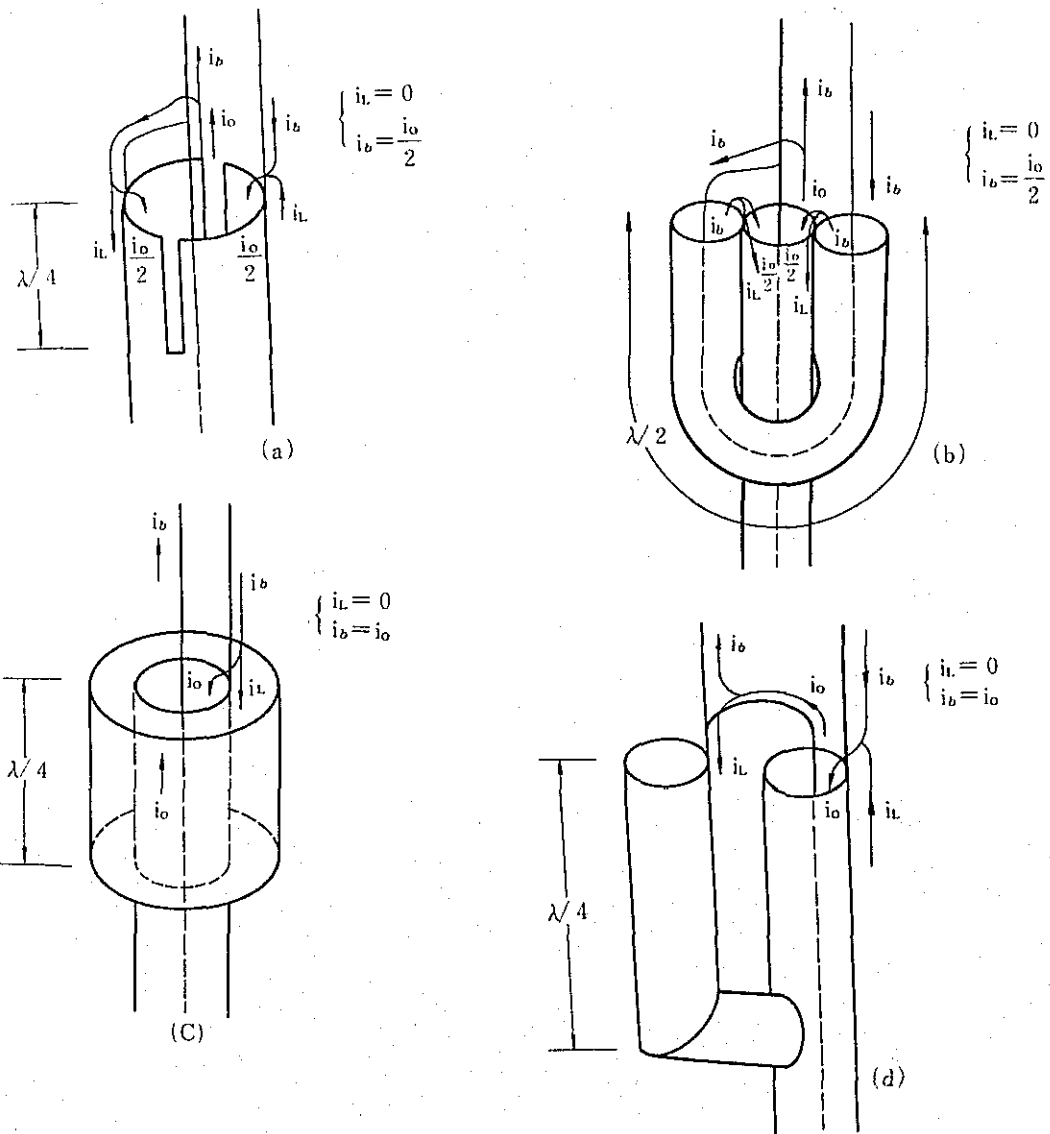


Figure 22.

