C.O.H. Reprints – N.S 165

Components of electrically small loop antennae: Part 2

In this section, *B A Bowles, B A Austin and C W P Attwell* describe the properties of matching networks employed which determine the most desirable network.

In Part 1 of this article it was shown that in the frequency range 100 kHz to 3 MHz, a typical electrically small loop antenna is equivalent to an inductance in series with a loss resistance, the two components being in parallel with a self-capacitance. It was also pointed out that in order to obtain an efficient transmission system a matching network is necessary to cancel the reactive antenna components and hence maximise the antenna current in the transmission mode and the output voltage in the receive mode.

This part of the article describes the properties of the matching networks that are frequently used with electrically small loop antennas and thereby determines the most desirable network.

Most matching networks use the simple two element "L circuit" which can transform a large range of load impedances to a real input impedance.

The eight forms of this circuit consist of various combinations of inductors and capacitors and have been examined by P. H. Smith,[1] who presented a series of graphs to enable the components to be selected. However, G. S. Smith,[2] has shown that the most efficient coupling networks for an inductive antenna consist only of capacitors and a network which incorporates all the various capacitive matching circuits is shown in Fig. 1.

Three cases of the network in Fig. 1 are of importance; these are:

(A) $C_1 = 0$
(B) $C_3 = 0$
(C) $C_1 = C_3 = 0$

The matching networks corresponding to the above are denoted by (A), (B) and (C) and it is evident that (A) and (B) are two of the eight forms of "L circuits". Network (C) has not been considered by P. H. Smith but it will be instructive to analyse this series tuned circuit since it has many advantages. Generally, a matching transformer is used with

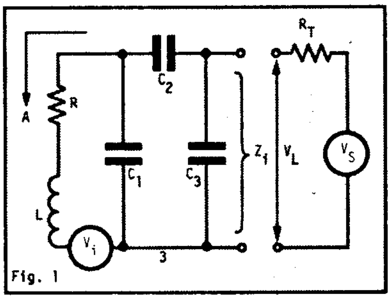


Fig. 1

Fig. 1: Universal capacitive matching network

(C) since its input impedance at resonance is low. However, the transformer will be omitted for the present calculations since its only effect is to change the magnitudes of the source impedance and EMF to the equivalent values shown in Fig. 1.

When the antenna is in the receiving mode, R_T is the load impedance (R_L) "seen" by the matching network, V_i is the induced voltage and V_s is zero. In the transmit mode V_i is zero and R_T is equal to the source impedance (R_S).

The most important parameters in a matching network are the magnitude and phase of the antenna current (I_A), the loaded output voltage (V_L) and the input impedance (Z_i). Each of these can be expressed by the following equations:

Graphs which enable the matching components to be selected when the magnitude of the antenna impedance is of the same order as the terminal impedance have been presented previously by P. H. Smith.[1] However, Part 1 of this article shows that the antenna resistance is a small fraction of the required terminal impedance and hence other methods of selection are necessary.

For matching network (A), the component selection procedure is done in two steps. The first step requires the antenna loss (R), antenna inductance (L) and antenna capacitance (C_1) to be known together with the required output impedance at resonance (f_0). The network output impedance at (f_0) can be derived from equation (3), which becomes:

$$\frac{(2\pi f_0 L)^2}{R}\left(\frac{C_2}{(C_2 + C_3)}\right)^2 + \frac{1}{j2\pi f_0 C} \qquad (4)$$

where $C = C_3\left[1 + \frac{C_1 C_2}{C_3(C_1 + C_2)}\right]$

For many applications, the reactive term can be ignored and the output impedance becomes:

$$\left\{\frac{(2\pi f_0 L)^2}{R}\right\}\left(\frac{C_2}{(C_2 + C_3)}\right) = R_0 \qquad (5)$$

The second step in the selection of C_2 and C_3 makes use of the reso-

$$\frac{I_A}{V_s} = \frac{Z_3 + jZ_4}{\{[Z_1 + R_s(Z_3 - Y_1)] + j[(Z_2 + R_s(Z_4 + Y_2)]\}\{Z_6 + jZ_7\}} \qquad (1)$$

$$\frac{V_L}{V_i} = \left\{\frac{Y_3}{Y_4 + jY_5}\right\}\left\{\frac{[Z_3 - Y_1 + jR_L(Z_4 + Y_2)]}{[R_L(Z_3 \times Y_1) + Z_1] + j[R_L(Z_4 + Y_2) + Z_2]}\right\} \qquad (2)$$

$$Z_i = \frac{Z_1 + jZ_2}{(Z_3 - Y_1) + j(Z_4 + Y_2)} = R_i + jX_i \qquad (3)$$

where $Z_1 = 1 - \omega^2 L\,(C_1 + C_2)$ $\quad Z_2 = \omega R(C_1 + C_2)$ $\quad Z_3 = -\omega^2 C_1 C_2 R$

$Z_4 = \omega C_2(1 - \omega^2 L C_1)$ $\quad Y_1 = \omega C_3 Z_2$ $\quad Y_2 = \omega C_3 Z_1$

$Y_4 = 1 - \omega^2 L C_4$ $\quad Y_5 = \omega C_4 R$ $\quad Y_3 = \frac{C_2}{C_2 + C_3}$

$C_4 = C_1 + \frac{C_2 C_3}{C_2 + C_3}$ $\quad \omega = 2\pi f$

The authors are with the Chamber of Mines of South Africa.

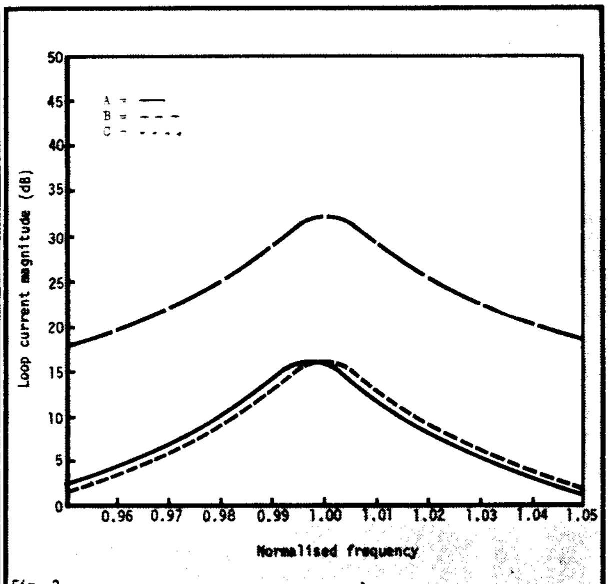


Fig. 2

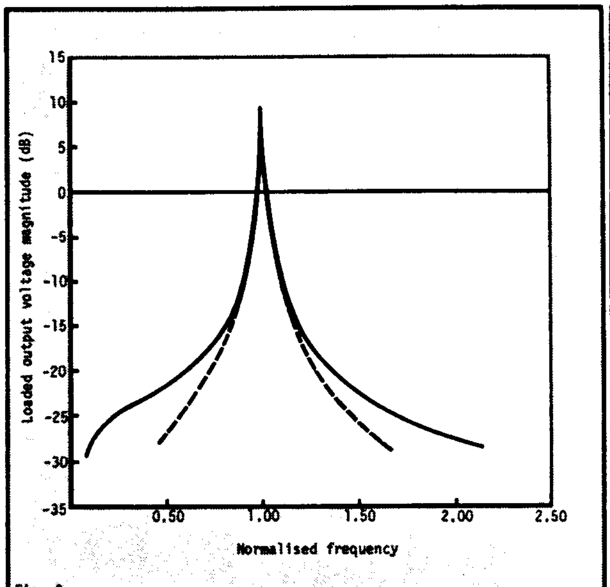


Fig. 3

nance equation, that is,

$$[(2\pi f_0)^2 L]^{-1} = C_1 + \frac{C_2C_3}{C_2 + C_3} = C_r \quad (6)$$

Hence the values of C_2 and C_3 can be found by solving equations (5) and (6).

The component selection for matching network (B) uses the variation of the effective inductance and loss of a parallel tuned circuit as shown in Part 1. As the frequency approaches resonance, both L_e and R_e increase. The component values are selected such at R_e equals the desired output resistance and C_2 is chosen to resonate the resulting value of L_e. The selection of C_1 and C_2 is therefore done in three steps. The first step requires R, L, f_0 and R_0 to be known and hence C_1 is solved using:

$$R_0 = \frac{R}{[1 - (2\pi f_0)^2 LC_1]^2 + (2\pi f_0 C_1 R)^2} \quad (7)$$

The value for C_1 is easily calculated by using a computer to plot R_0 against C_1. It should be noted that C_1 includes the antenna self capacitance.

The second step evaluates L_e from: (8)

$$L_e = \frac{L[1 - (2\pi f_0)^2 LC_1] - C_1R^2}{[1 - (2\pi f_0)^2 LC_1]^2 + (2\pi f_0 C_1 R)^2}$$

The third part resonates L_e with C_2 and hence:

$$C_2 = \frac{1}{(2\pi f_0)^2 L_e} \quad (9)$$

The selection of the components for matching network (C) requires the effective antenna inductance (L_e) and loss (R_e) to be known. Whereupon, C_2 is chosen to resonate with (L_e) and the value for (R_e) determines the turns ratio for the matching transformer.

Fig. 2: Variation: antenna current v frequency
Fig. 3: Loaded output voltage over a wide frequency range

Basic properties

In this section, the magnitude and phase of the antenna current, loaded output voltage and input impedance are compared for the three matching networks under consideration to establish the most suitable network.

The procedures outlined in the previous sections were used to calculate C_1, C_2 and C_3, for each of the matching networks to resonate an antenna at 913 kHz.

The output impedance required from (A) and (B) is 50 Ω and from (C) 1.23 Ω, which is equal to the loss resistance (R) of a particular antenna having an inductance of 21.4 μH: the antenna is assumed to have no self capacitance. Hence the antenna Q factor $\frac{\omega L}{R}$ is 100.

(A) $C_1 = 0$; $C_2 = 1{,}52$ nF; $C_3 = 22{,}30$ nF.

(B) $C_1 = 1{,}20$ nF; $C_2 = 221$ pF; $C_3 = 0$.

(C) $C_1 = 0$; $C_2 = 1{,}419$ nF; $C_3 = 0$.

The variation of the normalised antenna current with normalised frequency ($\frac{f}{fo}$ = 913 kHz) for the three networks when driven by matched source impedances is shown in Fig. 2. The normalised antenna current, expressed in decibels, is given by:

$$20 \log\left(\frac{100\, I_A}{V_s}\right) \text{dB}$$

The curves for networks (B) and (C) have identical shapes although (C) is displaced 16 dB above (B). This result is not surprising since no matching transformer is used with (C) which has an input resistance (R_i) at resonance of 1.23 Ω. If all the input power is dissipated in the antenna resistance (R), and Z_i at resonance is purely resistive then:

$$\frac{V_s^2}{4R_i} = I_A^2 R$$

Hence for identical antennas and voltage sources the ratio of the resonant currents in (B) and (C) are:

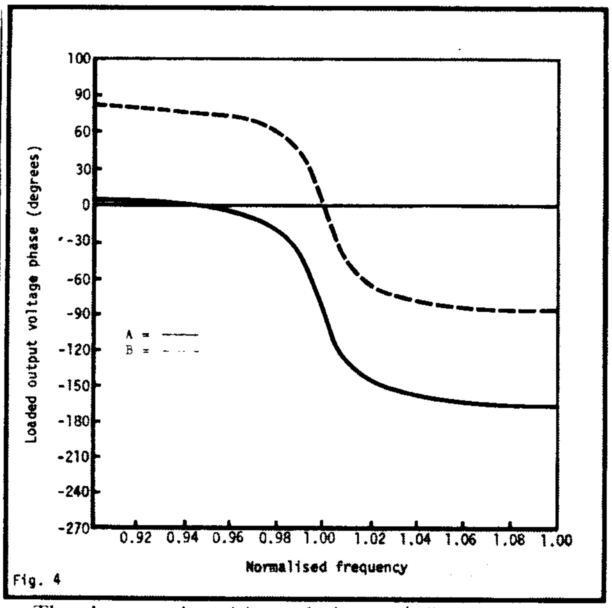
$$\left(\frac{R_i\,(B)}{R_i\,(C)}\right)^{\frac{1}{2}}$$

which expressed logarithmically is equal to 16.1 dB; which is the difference between the two responses in Fig. 2.

Network (C) with a perfect matching transformer therefore produces an equivalent maximum antenna current as in (A) and (B). However, the frequency response using network (A) is not quite the same, being displaced to a lower frequency range. A factor common to all curves is that the frequency difference at the 3 dB points ($f_2 - f_1$) is given by half the antenna Q factor, i.e.

$$f_2 - f_1 = \frac{2f_0}{Q}$$

The above relationship holds only for a source impedance equal to the matching network input impedance at the desired resonant frequency. The effect of other values of source impedance is treated later in this article.



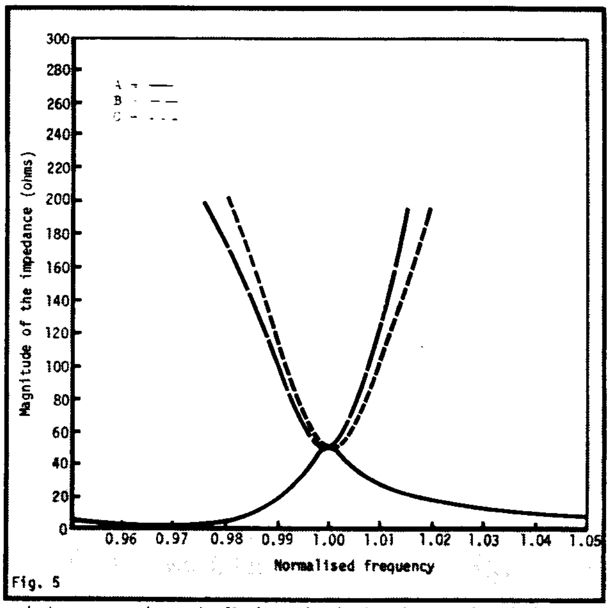


Fig. 4: Variation of loaded output voltage phase with frequency

Fig. 5: Vational: matching network input impedance with frequency

The shape and position relative to the frequency axis of the normalised output voltage frequency characteristics for networks (B) and (C) are identical to those in Fig. 2 if $R_L = R_s$. The shape of the characteristic for network (C) is also the same as that in Fig. 2 but the maximum output is 16 dB less than (B) or (C). This is due to the difference of terminal impedances, since no matching transformer is used for the calculations. Again, the response from (A) is shifted to a lower frequency range than either (B) or (C) and the bandwidth of all three is determined by Q/2.

Electrically small antennas therefore exhibit the property that the system gain and selectivity can be controlled by means of the antenna Q factor.

The effect of Q on the system output can be deduced easily if it is assumed that each network transfers the same input power to the load. Using these considerations then I_A and V_L are each proportional to $Q^{\frac{1}{2}}$ and for identical receiving and transmitting antennas the total system voltage gain is proportional to Q.

Frequency response of (A) and (B) over a wide frequency range is shown in Fig. 3. The response from (C) is not shown since it is identical to (B). It is apparent that (A) exhibits more outband attenuation than (B) above the resonant frequency but below the resonant frequency (B) is superior. For receiver applications these points are important since the noise performance of the system is influenced by the matching network selectivity.

The ideal frequency responses obtained for I_A and V_L imply that I_A and V_L are the reciprocals of each other when the terminal impedances are the same. Complete reciprocity also requires the phase responses to be identical and this has been found to be true. The phase response for (A) and (B) are shown in Fig. 4. The phase response of (B) is identical to that of (C). The phase response for all three networks is very linear over the system bandwidth which indicates that these antennas should give good performance in digital transmission systems.

Input impedance

The variation of Z_i with frequency for (A) and (B) and for (C) with a perfect matching transformer is shown in Fig. 5; the phase responses are shown in Fig. 6. It is apparent that (B) and (C) are similar since the impedance increases for frequencies away from resonance while the impedance for (A) decreases. An extended plot reveals that although (B) and (C) are similar within the system bandwidth, the difference between the two increases as the frequency deviates further from resonance. The phase responses for (B) and (C) are also different, both being capacitive below resonance, with (C) becoming inductive above resonance while (B) is at first inductive and then capacitive. Network (A) behaves in a similar manner to (B) but the inductive region is below resonance.

It is important to bear in mind that the impedance curves in Fig. 5 and Fig. 6 refer to one set of antenna specifications and that different results can be obtained for other antenna systems. In particular, it is not always possible to obtain a real value of input impedance and Smith[1] has shown that it is not possible to match all types of antenna impedance with networks (A) and (B). For example, if network (A) is designed for an output impedance of 5 Ω with the antenna described earlier, the phase response is shifted such that the peak just reaches zero for Q = 100 at a normalised frequency of 0.995. For all other frequencies the input impedance is capacitive. If the output impedance is 50 Ω, the critical Q factor is 10. In these cases, the capacitive term in equation (4) has a significant effect. If the Q factor for this antenna is reduced from 100 for an output impedance of 5 Ω the impedance is always capacitive and additional components are required to produce a real input impedance and hence obtain efficient power transfer.

A restriction of the range of antenna impedance than can be matched is also imposed by network (B). However, network (C) can always provide a real input impedance and furthermore, its impedance increases away from the resonant frequency. This is a desirable property for transmitter applications since a large part of the signal spectrum can be at the extremes of the system bandwidth and the supply current will be less for an impedance

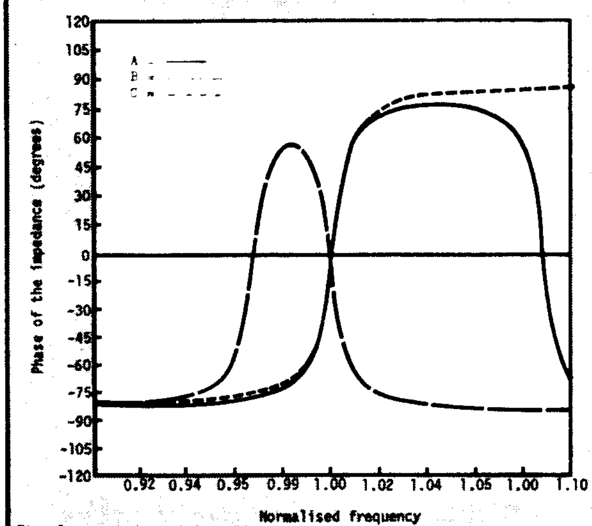


Fig. 6

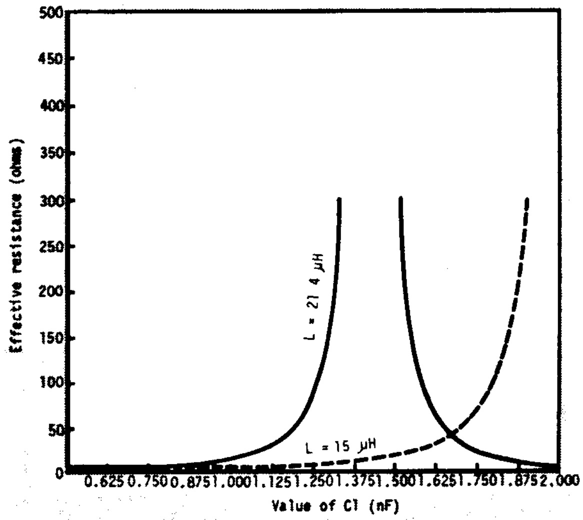


Fig. 7

Fig. 6: Variation: phase of matching network input impedance with frequency
Fig. 7: Effects of C_1 on effective resistance of a parallel tuned circuit

characteristic which increases rather than decreases. Network (A) however, exhibits a 60% decrease in impedance at the lower edge of the band with a consequent increase in current consumption.

The asymmetric decrease of impedance about resonance for network (A) is due to the series resonance of L and C_2 which can cause the input impedance to decrease to a very low value within the system bandwidth. The rate of decrease increases as the ratio $\frac{C_2}{C_3}$ decreases, that is for high inductance antennas or networks with a low output impedance. With some antenna systems a decrease of 80% at the lower 3 dB point has been calculated. The change of Z_i with frequency can cause a problem with portable equipment such as speech transceivers if variations in the matching or antenna components cause the resonant frequency to increase by greater than half the system bandwidth. In this situation the operating frequency could coincide with the impedance minimum and the operation of the radio transmitter and the RF filter of the radio receiver may be adversely affected.

A close examination of Fig. 2 and Fig. 5 reveals that the impedance peak for network (A) is at a higher frequency than the peak of the loop current. As the antenna Q factor reduces, the frequency separation between the two peaks increases and, with the antenna used for this example, if the Q factor is less than 20, the loop current at the frequency of the impedance maximum is less than the resonant peak by more than 2 dB. For an output impedance of 5 Ω the critical Q factor is 50. Consequently the output impedance peak is not an indication of resonance and if used for tuning the antenna it can cause misleading results. Conversely the output impedance may be significantly less at resonance than the maximum value and hence overload the transmitter.

These factors are important for intrinsically safe equipment since in these applications the antenna voltage is limited and a low impedance is often used to obtain higher transmission powers.

Component tolerances

The previous sections show that a high antenna Q factor leads to a higher input signal at the radio receiver and to better selectivity. However, before a value for the Q factor can be determined it is necessary to consider the sensitivity of the system parameters to practical variations in the components of the antenna and matching network.

The maximum change of frequency response allowable depends on the type of transmission system. If the modulation bandwidth is much less than the transmission bandwidth then the maximum change of resonant frequency is generally taken as half the transmission bandwidth. If the two bandwidths are comparable then the resonant frequency drift must be much less and depends on the type of modulation. For example, the minimum transmission bandwidth for a digital system is comparable to the baud rate.[3] In this case the frequency stability should be better than one tenth[4] of the data rate and for a resonant frequency tolerance of ±0.5% the antenna Q factor must be less than 40. For the purposes of this article, the component tolerances are considered for a system whose modulation bandwidth is much less than the transmission bandwidth.

Values of the matching network components, however, change with temperature, the environment and age by amounts which depend on the type of components used. It was pointed out in Part 1 that silver mica capacitors are recommended for receiver applications and polypropylene capacitors for transmitter matching networks. Silver mica capacitors have a long-term stability of about 0.1% and a temperature coefficient of ±50 ppm per °C. Hence a typical design tolerance for these components is ±0.2%. Polypropylene is not as stable (0.5%) as silver mica and has a higher temperature coefficient (−200 ppm per °C), and hence a typical design tolerance for these components is about ±1%.

The effect of capacitance variations on matching network (A) depends on the relative magnitudes of the components. For most applications the antenna impedance at resonance is much greater than the network output impedance and hence C_3 is significantly greater than C_2. Thus the resonance condition will be influenced primarily by C_2.

For reasons mentioned previously, this network is more suitable for receiver applications, in which case silver mica capacitors would be used. Assuming a tolerance of ±0.2%, the resonant frequency changes by ±0.1% (since $f_0 \alpha C^{-\frac{1}{2}}$) which implies that for this change in component value to be serious the antenna Q factor would be in the region of 1000, an impractical situation. Furthermore, the output impedance of the network is proportional to $\frac{(C_2)^2}{C_3}$ and hence a worst case error of ±0.8% is expected and this is also generally insignificant.

With network (B), however, the component sensitivity is not quite so clear since the parallel capacitance (C_1) has a complex effect on L_e and R_e. If R_e is plotted against C_1 at 913 kHz then Fig. 7 shows that large values of R_e are very sensitive to C_1. Even when $R_e = 50\,\Omega$ the effective resistance changes by about 10% for a 1% change of C_1. The change of L_e is not quite so dramatic, being about half the change of R_e and hence for a 1% change of C_1, the series resonant frequency will change by about 2.5%. Hence the Q factor should be less than 40 in order to achieve a stable system.

The effect of C_2 using network (C) is straightforward. In practice this network is suited to transmitter applications and hence the maximum change of resonant frequency with polypropylene capacitors should not exceed 0.5% and the Q factor must be less than 200.

Inductance effects

The antenna inductance is hardly affected by changes in temperature. However, its value is affected by other factors, such as the presence of conducting objects and by physical movement if the antenna is flexible and is moved. Measurements at 200 kHz indicate that the inductance can be reduced by 7% when indirectly coupled within a few centimetres to underground power-line cables. The inductance of an antenna worn across the chest may change by ±2% due to movement of the body.

If the equipment is portable then the maximum change of frequency response must lie within the transmission bandwidth. With network (A), the antenna Q factor should be less than 100 for the majority of mining applications and the transmitter must be able to deliver power into the reduced matching network input impedance as the impedance response drifts. The antenna Q factor using network (C) should also be less than 100 to avoid retuning but although the transmitter will generally not be harmed by a load impedance higher than nominal it would limit the maximum output voltage with a consequent increase in transmitted harmonics. The effect of an inductance change using network (B) is far more serious and Fig. 7 indicates that for a 30% reduction of L, R_e falls from 50 Ω to 6 Ω. For a 2% reduction of L, both R_e and L_e reduce by almost a half and complete retuning is necessary.

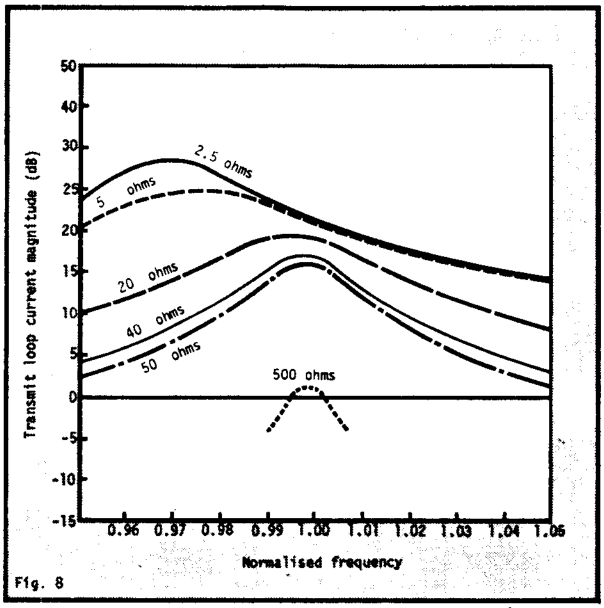


Fig. 8: Effect of source resistance on loop current frequency response using network (A)

In some permanent installations, such as the transmission of telemetry data in shafts, indirect coupling into power-line cables is used to achieve a considerable increase in received signal strength. The received signal can be increased further by tuning the antenna in-situ and practical measurements indicate that 100% improvement can be obtained. Using network (A), C_2 is adjusted to bring the resonant peak for either I_A or V_L coincident with the frequency of operation. Using this procedure Z_0 does not change since its real part reduces to

$$\frac{1}{\omega^2 C_3{}^2 R}$$

However, with L smaller than the free space value, the antenna Q factor is reduced and the transmission bandwidth is increased. Retuning using network (C) is also simple but with network (B) both matching components must be adjusted by means of a fairly complex procedure.

The effect of source impedance (R_s) on the antenna current (I_A) using network (A) is shown in Fig. 8. As the source impedance decreases from 500 Ω, the bandwidth widens with the response at first being centred just below 913 kHz and then moving to lower frequencies. For values of source impedance less than 2 Ω, the current maximum occurs at the series resonant frequency of the network. The bandwidth increases until the source impedance is about 5 Ω, whereafter it decreases. When $R_s = 5\,\Omega$ the bandwidth is double the value for a matched system. The real and imaginary parts of the antenna input impedance at the maximum value are given in Table 1.

From this information it is possible to calculate, for a lossless matching network, how much extra reactive power must be supplied for the same power into R_{IA} and hence the same antenna current. The calculated values are given in Table 2.

Alternatively, the increase in bandwidth could have been obtained with $R_s = 50\,\Omega$ by decreasing the antenna Q factor. If the bandwidth is to be doubled, the Q factor must be halved and hence the antenna current will fall by $Q^{\frac{1}{2}}$ or 3 dB. Therefore, if the same value of peak current is to be maintained the antenna input power must be increased by 3 dB which is greater than the value of 1.92 dB required by using a source resistance of 5 Ω. Therefore, a mismatched source impedance can be used with this network to achieve a greater transmission bandwidth. However, it is considered that there is little advantage operating in this mode since the network is effectively equivalent to network (C) with the disadvantage of a reactive input impedance and hence poorer power transfer.

Since these matching networks are reciprocal networks, the same frequency responses are obtained in the receiving mode. However, it may happen that the transmitter output impedance is not the same as the receiver input impedance and Fig. 8 shows that the frequency response in the two modes will not be the same. If the antenna has the same parameters as those used for Fig. 8, little trouble should be experienced for voice communications providing the transmitter output impedance is kept between 20 Ω and 50 Ω and the receiver input impedance is 50 Ω. For the transmission of data pulses, over a narrow band transmission link, the tolerances on the impedances should be tighter since the pulse waveform is sensitive to the variation of amplitude and phase with frequency and the performance of the rest of the system may be critically affected.

With some transmitter circuits, the output impedance can be adjusted independently of the output power. If the output impedance of such a transmitter is increased above 50 Ω while keeping the output power constant the system bandwidth reduces but the maximum antenna current remains the same. This is not a desirable mode of operation since the reduced bandwidth could be obtained with a higher antenna Q factor with a consequent increase in the maximum antenna current. It is therefore concluded that the optimum source impedance for network (A) is equal to the network input impedance.

The effect of source impedance variations on networks (B) and (C) is to change only the system Q factor; the resonant peak remains at the same frequency. As the source impedance is reduced the responses become narrower and hence the effective antenna Q factor increases. For an antenna of a defined number of turns this property can be used to advantage to increase the antenna current[5] while keeping the system Q factor constant. In order to achieve this it is necessary to have the greater part of the total series resistance which determines the system Q factor in the output impedance of the transmitter where by suitable circuitry[6] it absorbs little power. As the antenna loss (R) decreases the transmitter output impedance must increase to keep the system Q factor constant, but for a transmitter of defined output power (P_0) the antenna current is given by:

$$I = \left(\frac{P_0}{R}\right)^{\frac{1}{2}}$$

Hence by using a low loss antenna and a matching transformer with a high turns ratio it is possible to obtain a high value of antenna current. An antenna using these principles has been built and has produced a 200μV (p.d.) signal at a receiver antenna when indirectly coupled to 2.5 km of 6.6 kV power cable. The transmitter had an output power of 10 W.

$R_s(\Omega)$	$R_{IA}(\Omega)$	$X_{IA}(\Omega)$
2.5	1.41	0.51
5.0	2.44	2.91
20	31.18	18
50	51	0
500	51	0

Table 1: Real and imaginary parts of antenna input impedance at maximum value

$R_s(\Omega)$	Extra transmitter power (dB)
2.5	0.27
5	1.92
20	0.62
50	0
500	0

Table 2: Calculated values of extra reactive power required

Network (A) is used frequently for matching loop antennas since it is simple to design and being only capacitive is highly efficient. However, for transmitters it has the disadvantage of a sharp reduction of input impedance on the low-frequency side of the resonance curve and thereby increases the consumption of current by the transmitter, if the modulation bandwidth is comparable to the transmission bandwidth. In radio receiving systems it has the disadvantage of lower outband attenuation compared to the other matching networks for frequencies below resonance which is a region that generally has a greater noise content. In situations where the antenna Q factor is low it is difficult to provide a terminal impedance which is real and which has a low value. The optimum source impedance is equal to the input impedance of the network and its value should be tightly controlled if a real input impedance is required at the system resonant frequency.

Matching network (B) produces an identical frequency response to network (C). However, network (B) is restricted in cases of antenna detuning since both L_e and R_e are sensitive to capacitance and inductance variations. Typical long-term stability and temperature variations of the matching components restrict the antenna Q factor to less than 40. If retuning is necessary then both matching components must be adjusted which is a difficult task in the field. However, this network has an increasing impedance characteristic away from resonance which is a desirable characteristic for transmitter applications.

The problems existing with matching networks (A) and (B) are not present with (C). Since only one resonating component is used the task of retuning is much simpler than with (A) or (B). The impedance characteristic increases at frequencies away from resonance and its lower outband attenuation is better that that of (A). Furthermore, it is theoretically possible to match the antenna to any real value of terminal impedance. However, it has the disadvantage of requiring a transformer to match the low-impedance antenna to a transmission line and for a high turns-ratio the design of a low-loss transformer is no easy task.

Minimum transmission bandwidth is determined partly by the effect of the mine environment on the maximum change of all the components of the antenna system. It is also influenced by the distortion which may be introduced into the demodulated radio frequency signal and the decreasing noise performance of some modulation methods at low bandwidths. The increase of RF signal to noise ratio with antenna Q factor poses some interesting problems for the selection of a radio modulation method.

References

1. Smith, PL H. L-type impedance transforming circuits. Electronics, Vol 15, No 127, pp 48-54, March 1942.
2. Smith, G. S. Radiation efficiency of electrically small multi-turn loop antennas. IEEE Transceivers of Antennas and Propagation. Vol AP-20, pp 656-657, September 1972.
3. Sunde, E. D. Ideal binary pulse transmission by AM and FM. The Bell System Technical Journal, pp 1357-1426, Vol 38, Nov 1959.
4. Thompson, R. and Clouting, D. R. Digital angle modulation. Wireless World, pp 69-72, Feb 1977.
5. Foden, R. Private communication within The Chamber of Mines.
6. Bowles, B. A. Full duplex transmission for use underground. Electronic Engineering, Vol 50, No 604, pp 42-43, April 1978. □