AMPLIFICATORE DI SERVIZIO – KOSS

RIFERIMENTI

Genere	DATA	Generalità	Note	Distribuzione
RADIO	APRILE 2020			AF WEB

GENERALITA'

Ho approfittato del fare pulizia in garage per recuperare una mini cassa audio per computer, modello KOSS, pensando di usarla come ampli audio di servizio. Ne ho già un'altra, che utilizza l'"ubiquitous" LM386, che è montata fissa sulla postazione di lavoro. Inoltre una simile, portatile con speaker esterno, che usavo per le lunghe trasferte in albergo.

Ho notato che questa cassa ha altoparlante da 3Watt molto robusto e subito mi ha invogliato ad usare un integrato più potente del 386, il TDA1517 di cui ne ho un paio regalatemi da Sauro IZ5GSF. Mi son chiesto perché li ha comprati se poi non li utilizza? boh!

È una delle prime cassette stereo, che avevano una reale se pur minima consistenza nei componenti interni, anche se l'amplificatore originale è piuttosto scarso e montato in una sola delle due cassette. Ho lavorato su quella senza amplificatore. Le serie più moderne di diffusori per PC hanno ridotto drasticamente le prestazioni a favore di design quanto più strampalati.

L'amplificatore di servizio, come i precedenti, incorpora un filtro passa basso per limitare la banda allo stretto necessario per la voce e anche meno.

Nella foto la piccola cassa, 9 X 13 X 9 cm, già modificata con manopola del volume e jack 3.5mm mono per l'ingresso segnale.

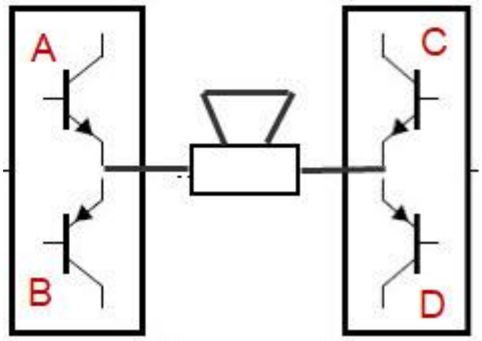
Rispetto alle altre versioni l'ingresso va direttamente ai capi di un potenziometro da 250 kOhm, in modo da essere ad "alta" impedenza.

"No frills" così niente interruttore del volume, solo la protezione da inversione polarità dell'alimentazione, che è quella di stazione, ovvero 11-15 Volt dc.

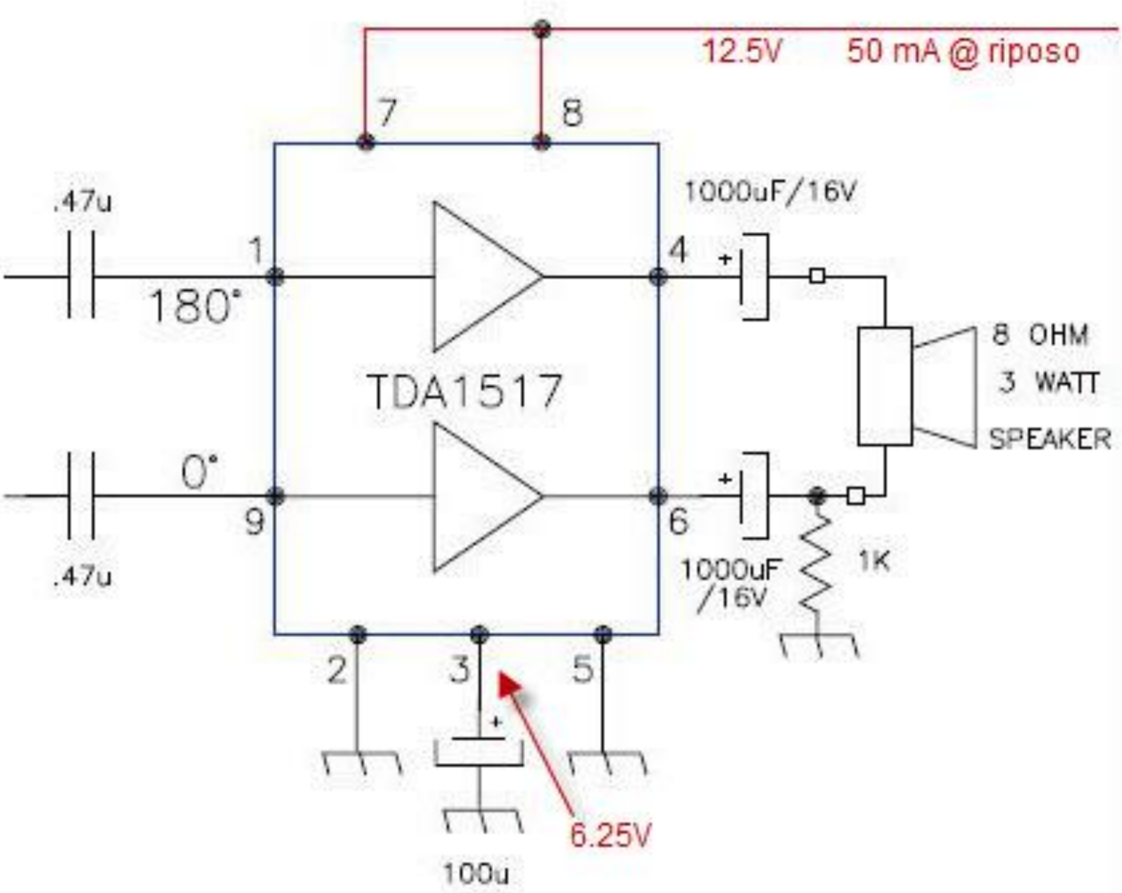
FINALE POTENZA

Come già detto il finale è basato sul circuito integrato TDA1517 che ha al suo interno due amplificatori da 6 W ciascuno. Normalmente viene impiegato per amplificatori stereo.

Io lo impiego come amplificatore mono ed un solo altoparlante, in una configurazione che è a H o a ponte, anche se non si evince facilmente dallo schema. La figura qui sotto è più esplicativa.



Quando conduce A conduce D, e alternativamente conducono B e C. Sul carico la tensione alternata è doppia di quella che si svilupperebbe in circuito single ended.



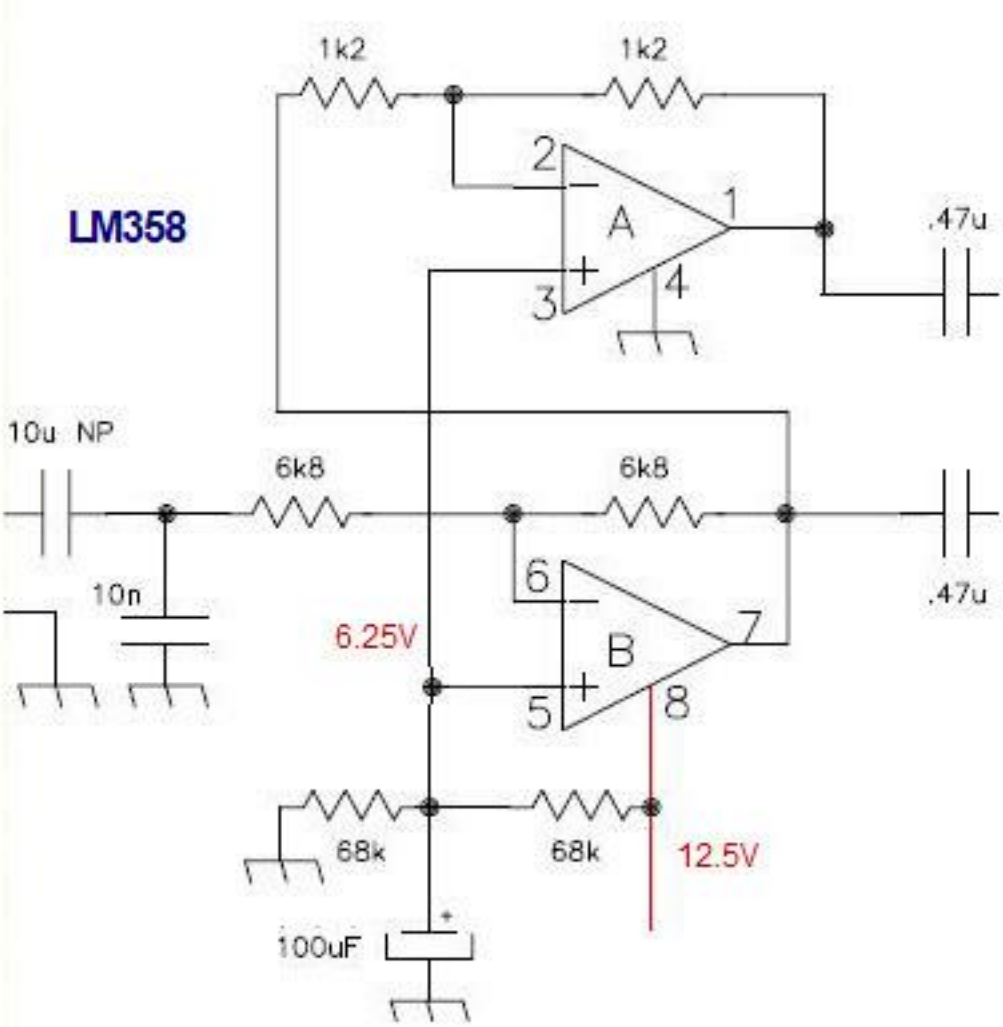
Nella pratica ho provato a raggiungere i limiti e misurando ai capi dell'altoparlante ho misurato 7 Volt rms a 1 kHz. Essi corrispondono a 20 Vpp, mentre l'alimentazione era a 12.5Vdc

Il TDA1517 non è risparmioso: a riposo senza segnale consuma 45 mA. Il costruttore ha previsto un pin, il pin 8, che collegato al + attiva l'amplificatore, se lasciato aperto mette in st-by con un consumo irrisorio. Lo ho lasciato sempre attivo, essendo un amplificatore di servizio che si utilizza in laboratorio.

Sul piedino 3 è presente la tensione interna di metà alimentazione, che pensavo di utilizzare insieme al circuito di pilotaggio per una configurazione minimalista senza condensatori ne di ingresso ne di uscita, ma poi non ho avuto voglia di sperimentare.

PILOTA SFASATORE

Fondamentalmente serve pilotare i due ingressi del finale con due segnali sfasati di 180° tra loro. Di questo si occupa l'operazionale A nello schema sotto. È un amplificatore a guadagno 1 con resistenze di polarizzazione abbastanza basse di valore per non creare fenomeni di filtro con le capacità presenti nel circuito.



Anche B è un amplificatore a guadagno unitario, usato come buffer per non avere il pilotaggio dell'ingresso 9 del finale dipendente dall'impedenza del pilotaggio.

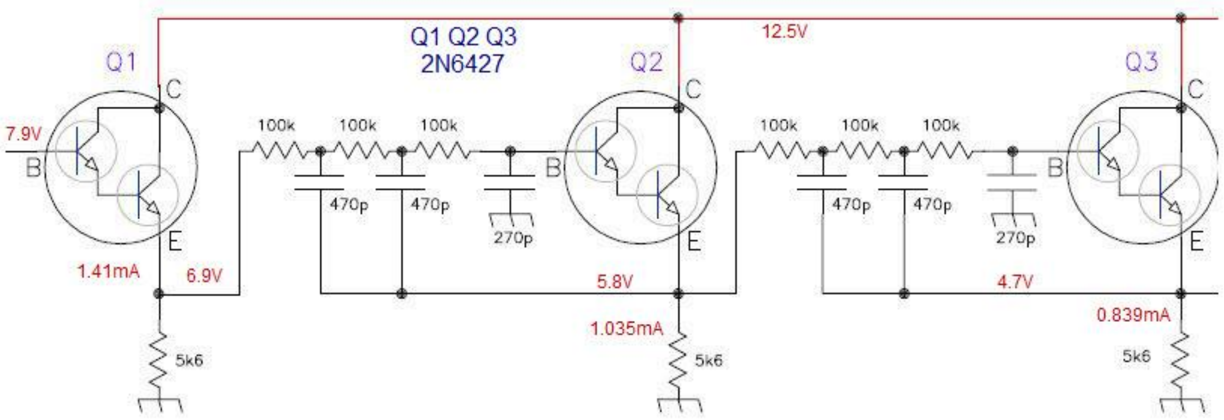
Il condensatore di ingresso è un elettrolitico tipo NP non polarizzato. C'è anche un condensatore da 10nF verso terra, l'ho messo perché alla prima accensione con il volume al massimo si sentiva RAI1, piano ma da far prendere provvedimenti.

FILTRO PASSA BASSO

In fondo a questa nota ho riportato l'estratto da un numero di VHF COMMUNICATION, del 1969 credo, a firma di Schmitzer alias DJ4BG, sulla realizzazione di semplici filtri attivi audio. DJ4BG era un mio idolo da ragazzo, anche se non appare molto teorico, però è pratico ed incisivo.

Ho tratto da quelle note uno schema e l'ho riprodotto quasi fedelmente.

I transistor sono dei darlington in unico contenitore TO92, non troppo comuni, acquisto TEKKNA, ma potrebbero essere anche dei più comuni BJT singoli collegati darlington.



Ho modificato la rete di polarizzazione di Q1 in modo da avere sugli emitter Q1 Q2 e Q3 dei valori di tensione che fossero centrati rispetto all'alimentazione. Tutti e tre gli stadi sono a guadagno unitario e l'effetto di picco vicino alla frequenza Fc è con i valori selezionati molto limitato.

CONCLUSIONI

Ancora nel circuito c'è un potenziometro subito all'ingresso e un diodo di protezione dall'inversione di polarità.

Ho messo un LED verde sotto alla griglia per avere idea dello stato di acceso dell'unità.

Le misure a 1 kHz danno 700 mVpp in ingresso per avere 3 Watt reali sull'altoparlante. 300 mVpp per mezzo Watt e 140 mVpp per 100 milliWatt in uscita.

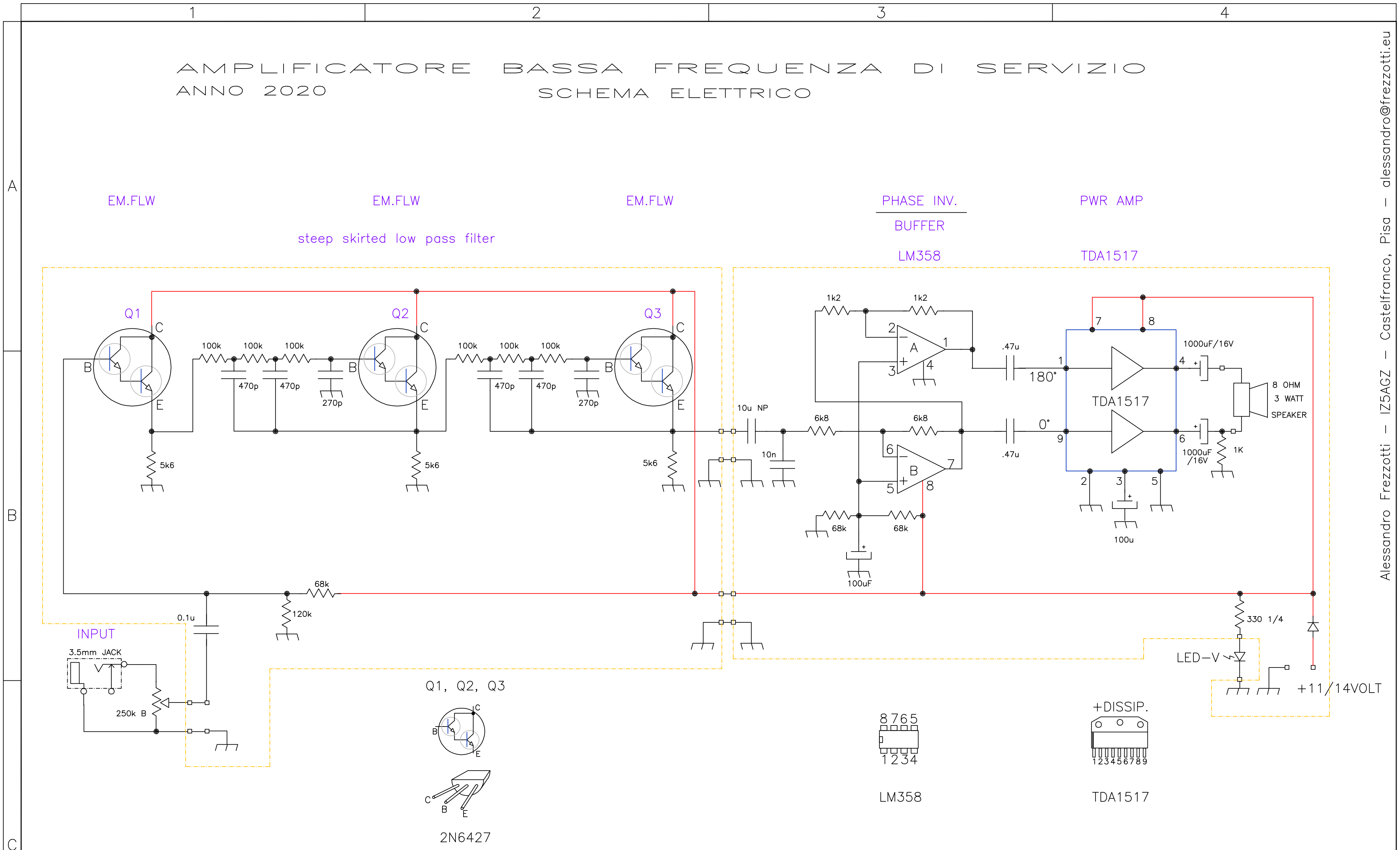
Esagerando, con 6 Watt in uscita il consumo di alimentazione totale si aggira sui 600-700 mA dal 12.5 Volt

Buon divertimento, Alessandro Frezzotti

AMPLIFICATORE BASSA FREQUENZA DI SERVIZIO

ANNO 2020

SCHEMA ELETTRICO



NOTA: pin7 TDA1517 alimentazione, pin 8 a +12 abilita uscita, se aperto output muto
DISEGNO NON IN SCALA
C IN uF DOVE NON INDICATO
R IN OHM DOVE NON INDICATO

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ACTIVE AUDIO FILTERS - PART I

by D.E. Schmitzer, DJ 4 BG

1. INTRODUCTION

Active audio filters are to be described which obtain favourable attenuation curves without using inductances. A principle is given that allows filters to be dimensioned according to prototype measurements. Due to the availability of integrated circuits, the author explains how operational amplifiers can be used for filter configurations which simultaneously offer a noticeable amplification. Practical circuits are not given; they are given in the second part of this description.

It is very advantageous to keep the bandwidth of radio equipment as narrow as possible. This is not only to satisfy often disregarded official requirements but also to keep required transmission bandwidth, and thus interference to other stations, at a minimum. It should be considered that any transmit bandwidth that is greater than that required, represents wasted transmit energy. On the receive side, an excessive bandwidth will cause a reduction of the signal-to-noise ratio and thus a reduction of the receiver efficiency.

It is possible, using some circuits, to build up filter configurations using only resistors and capacitors (RC combinations) instead of inductances and capacitors (LC combinations). Several types of active RC filters are to be described that are especially suitable for tailoring the voice frequency range.

2. CIRCUITS

The four most simple and easily understandable configurations of the many known active filter circuits have been chosen, from which only two will be considered in detail (1), (2).

2.1. BASIC LOW-PASS AND HIGH-PASS CIRCUITS

Figure 1a shows a high-pass filter using shunt resistors, and Fig. 1b the derived low-pass filter. High-pass and low-pass filters using series resistors are shown in Fig. 2a and 2b. Since inductances for filter applications in the audio frequency range are large, heavy and expensive components, only filter circuits based on Figures 1a and 2b, i.e. RC filters, will be explained further.

The amplifiers contained in the basic circuit diagrams are assumed to be ideal impedance transformers having a gain of A = 1, an infinitely high input impedance, an infinitely low output impedance and no phase shift. In practice, a sufficiently good approximation of such amplifiers can be achieved using a common-collector transistor configuration (emitter follower).

2.2. ATTENUATION RESPONSE

The attenuation curves obtainable with these circuits have a maximum skirt slope of 18 dB per octave or 60 dB per decade. The fundamental difference between a high-pass and a low-pass filter is: The attenuation of the high-pass filter using shunt resistors (Fig. 1a) strives towards infinity on decreasing

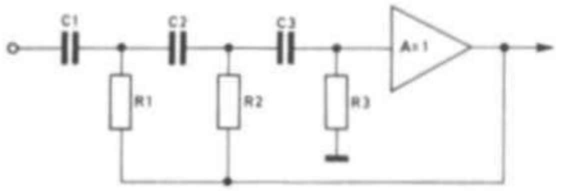

Fig. 1a High-pass filter with parallel resistors

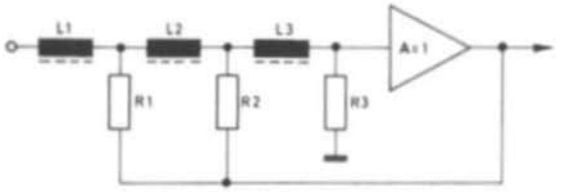

Fig. 1b Low-pass filter with parallel resistors

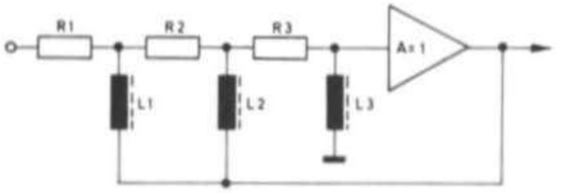

Fig. 2a High-pass filter with series resistors

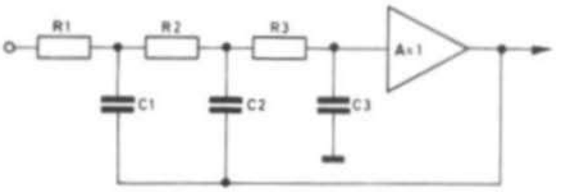

Fig. 2b Low-pass filter with series resistors

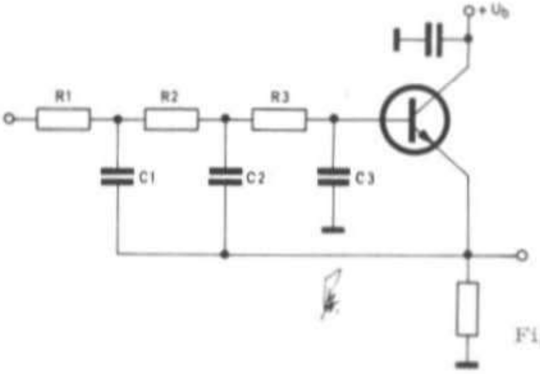

Fig. 3 Operation of the low-pass filter at high frequencies

the frequency, whereas the attenuation of the low-pass filter according to Fig. 2b approaches a finite value on increasing the frequency; this is because a voltage division between resistor R 1 and the output impedance of the amplifier will only occur at high frequencies. With a bipolar transistor in a common collector configuration, this output impedance Z_{out} is equal to the input impedance Z_{in} of the transistor in a common-base configuration since the last shunt capacitor C 3 shorts the base to ground at higher frequencies (Fig. 3). It is thus possible to calculate the maximum stop band attenuation of the low-pass filter whilst dimensioning the circuit since Z_{in} can be calculated according to equation 1a or 1b.

$$Z_{in} \approx \frac{1}{S} \approx \frac{26\ mV}{I_e} \approx \frac{26\ mV}{I_c}$$

Equation 1a: Input impedance of a bipolar transistor in a common-base configuration. Where S = transconductance, I_e = emitter current, I_c = collector current.

A transistor having a collector current of 1 mA will have an input impedance of 27 Ω in a common-base configuration. If a resistor of 10 kΩ is selected for R 1, the maximum stop band attenuation a_{max} is:

$$a_{max} = \frac{10\ 000\ \Omega}{26\ \Omega} = 385 \mathrel{\hat{=}} 51.7\ dB$$

Higher attenuation values can be obtained by increasing the collector current so that the input impedance Z_{in} is decreased or by correspondingly increasing the value of resistor R 1. Both cases can cause difficulties because the DC operating point of the transistor must be adjusted and maintained. An improvement can be achieved by using a Darlington circuit according to Fig. 4 instead of a single transistor. The Darlington circuit allows greater series resistance values to be used since the increased current amplification means that less base current will flow for the same collector current.

When calculating the maximum possible attenuation obtainable with this configuration, it should be noted that the input impedance of the Darlington circuit is somewhat higher than that of a transistor having the same collector current. This is because the base of the second transistor in a Darlington circuit is not directly connected to zero potential in the AC sense but via the input impedance value of the first transistor as given in equation 1a. The collector current of the first transistor must therefore be included in the calculation. However, since this current is very low, the impedance, which is no longer negligible, will also appear at the output reduced by the current amplification factor β_2 of the second transistor.

$$Z_{in}\ (D) \approx \frac{26\ mV}{I_{c2}} + \frac{26\ mV\ B_2}{I_{c2}\ \ \beta_2}$$

Since the static current amplification B_2 and the dynamic current amplification β_2 are approximately equal, the following offers a good approximation

$$Z_{in}\ (D) \approx \frac{50\ mV}{I_{c2}}$$

Equation 1b: Input impedance of a Darlington circuit (Fig. 4).

Very large series resistors are permissible when a field effect transistor is used in a common drain amplifier circuit as given in Fig. 5. This configuration leads to high stop band attenuation values although the output impedance is one or two orders-of-magnitude higher than that of bipolar transistors due to the low transconductance ($Z_{out} \approx 1/S$, is in the same order as for vacuum tubes). However, the attenuation of approximately 60 dB attainable with a single transistor should be sufficient for most applications.

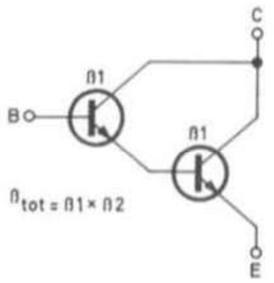

Fig. 4
Darlington Circuit

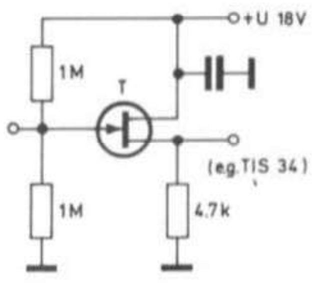

Fig. 5 Field effect transistor in common-drain circuit

2.3. PROTOTYPE MEASUREMENT AND MEANS OF CALCULATION

Since active filter circuits are not easily calculated, some fundamentals were laid down. A prototype measurement was made under these conditions from which any desired cutoff frequency and within limits, any curve form can be derived. The prototype measurement was made on a low-pass filter, as shown in Fig. 2b, according to the following considerations:

The series resistors R 1, R 2 and R 3 are of the same value, the shunt capacitors C 1 and C 2 are also equal and it is only the value of C 3 that is varied (For a high-pass filter as given in Fig. 1a, the following is valid: C 1 = C 2 = C 3, R 1 = R 2; R 3 is variable). By reducing the value of C 3 (or increasing R 3 for the high-pass filter) with respect to C 1 and C 2 (or R 1 and R 2 for the high-pass filter) it is possible for the transition between the pass band and stop band to be varied up to overshoot conditions; at lower values of C 3, the circuit can even break into oscillation, but this is not considered here. Since it has been found that the same principle is valid for both high-pass and low-pass filters, only the low-pass filter need be discussed.

The curves obtained from the prototype circuit according to Fig. 6, are shown in Fig. 7. To redimension the circuit for other cutoff frequencies, it is necessary to base same on the cutoff frequency of the first RC-link comprising R 1 and C 1. This frequency is designated f_1 in our example.

$$f_1 = \frac{1}{2\pi \; R1 \; C1} \qquad \text{Equation 2}$$

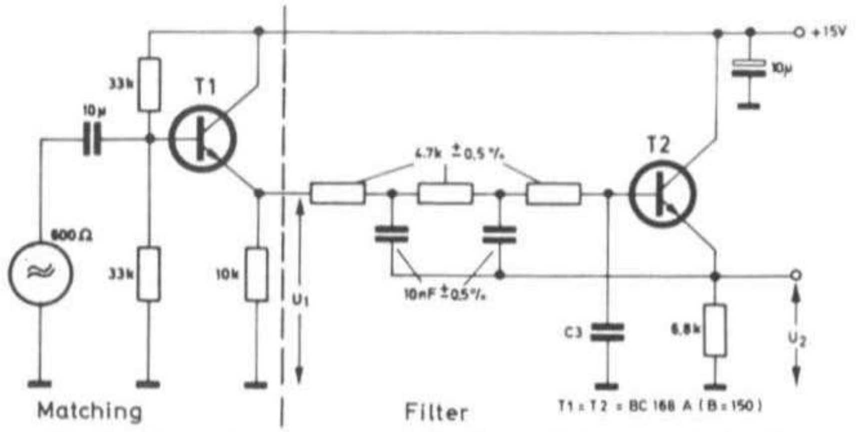

Fig. 6 Measuring arrangement for the prototype measurement

This frequency will only coincide to the -3 dB limit frequency when C 3 = 0.45 x C 1. The amplitude of the overshoot with respect to the level at frequencies well below the cutoff frequency is not considered during this definition. Figure 8 has therefore been provided as an additional aid to display the relationship between f_1 and $f_{-3\ dB}$ as a function of the relationship of C 3/C 1. If a filter for another frequency is to be recalculated from these two prototype curves, this should be made in the following manner: Firstly select the required attenuation curve from Fig. 7 which in turn lays down the relationship C 3/C 1. Fig. 8 then indicates the relationship $f_{-3\ dB}/f_1$.

Resistor R 1 or capacitor C 1 are given and the value of the appropriate second component, i.e. C 1 or R 1, must be determined with the aid of f_1 from equation 2. Capacitor C 3 is determined from C 1 and the relationship C 3/C 1 indicated in Fig. 7. Since it was determined that R 1 = R 2 = R 3 and C 1 = C 2, all frequency-determining components of the circuit are known.

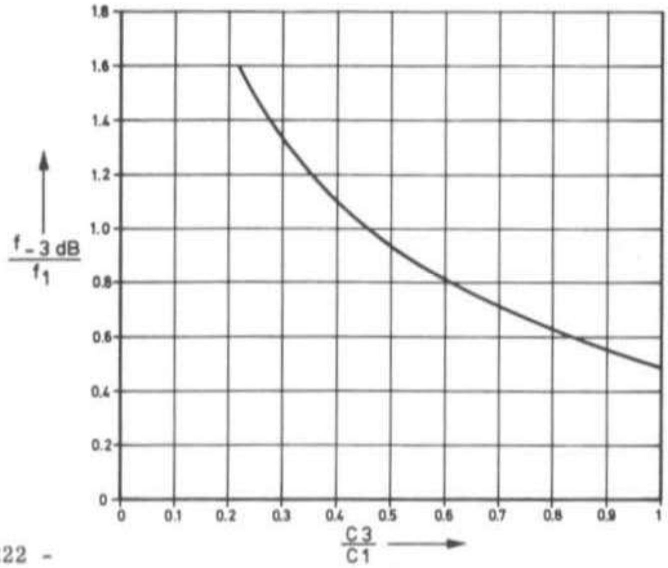

Fig. 8

Relationship between $\frac{C\ 3}{C\ 1}$ and $\frac{f\ -3\ dB}{f_1}$

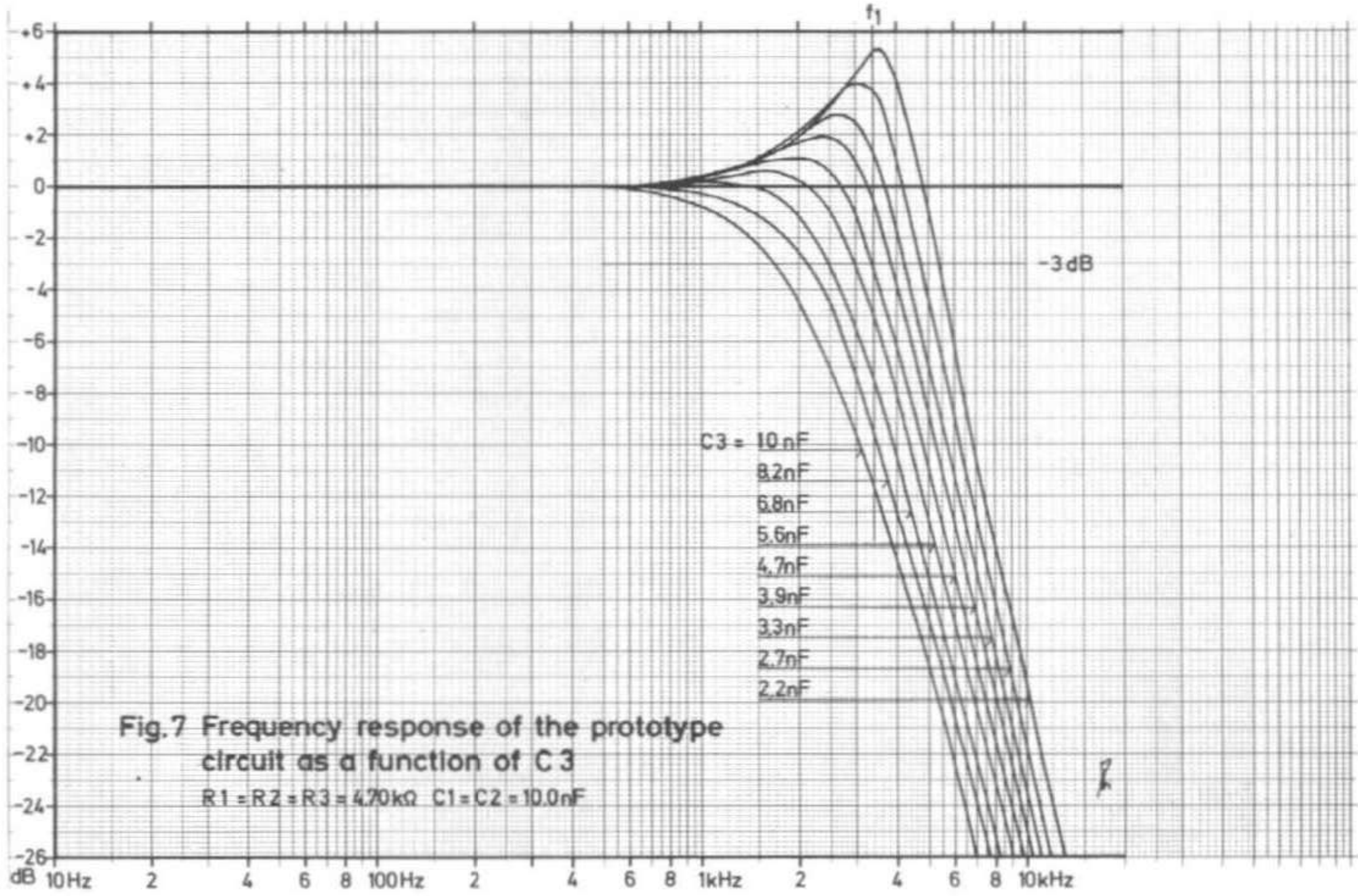

Fig.7 Frequency response of the prototype circuit as a function of C3

R1 = R2 = R3 = 4.70kΩ C1 = C2 = 10.0nF

2.3.1. EXAMPLE

A low-pass filter with the following specifications is required: $f_{-3\ dB} = 2.7$ kHz and an overshoot of max. 1 dB. Also given is: R 1 = R 2 = R 3 = 10 kΩ. Required are the values of C 1, C 2 and C 3. Figure 6 shows that C 3 was 5.6 nF in the prototype (C 1 = C 2 = 10 nF) in order to obtain an overshoot of approx. 0.6 dB. This indicates: that C 3/C 1 = 0.56 and that according to Fig. 8 $f_{-3\ dB}/f_1 = 0.88$; f_1 is thus 2.7 kHz/0.88 = 3.07 kHz.

Equation 2 shows that: C 1 = 5.18 nF and C 3/C 1 = 0.56 so that C 3 = 2.9 nF. Normally, conventional standard values must be used (e.g. C 1 = C 2 = 5 nF and C 3 = 2.7 nF) and the series resistors somewhat varied if this should be necessary when deviating greatly from the required value.

3. ACTIVE FILTERS USING OPERATIONAL AMPLIFIERS

An operational amplifier with negative feedback down to a voltage amplification of A = 1 is essentially better with respect to satisfying the demands of an ideal impedance transformer than is a simple transistor. Such amplifier configurations are available as integrated circuits at prices well within reach of amateurs. Since these components have very low input current requirements, they are very similar to the darlington circuit given in Section 3.2. which means that they are suitable for use with filters having rather high series resistance values and low shunt capacitance. This is especially true when an additional resistor (R 4) is used to compensate for the voltage drop across the series resistor (see Fig. 9). This resistor should have a value equal to the sum of the resistance values at the non-inverted input.

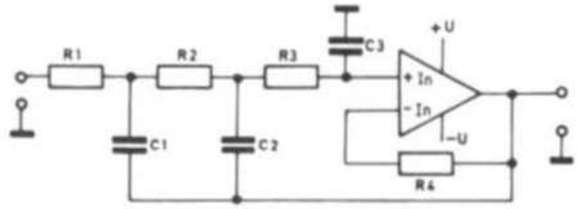

Fig. 9 Filter using an operational amplifier

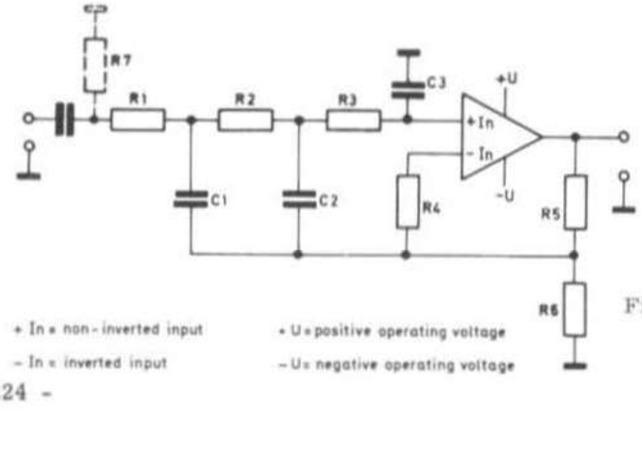

Fig. 10 Filter and amplifier using an operational amplifier

By slightly extending the circuit, it is possible to not only use it as an active filter, but also to obtain an additional amplification. Such a circuit is given in Fig. 10. If the remaining negative feedback is sufficient, the available amplification will only be dependent on the relationship (R 5 + R 6)/R 6. Since conventional operational amplifiers offer non-load amplification values of over A = 1000, it is possible to obtain operational amplification values of up to A = 100, corresponding to 40 dB. The resistance relationship in the negative feedback loop is thus 100 : 1, for instance, R 5 = 100 kΩ, R 6 = 1 kΩ.

Operational amplifiers are usually operated from two operating voltages, i. e. one positive and one negative voltage to ground (zero), for instance ± 6 V. The inputs should be connected to zero in the DC sense. If this occurs using an additional resistor (R 7 in our example), it will be necessary to increase the value of R 4 by the same amount. In actual fact, the following must be fulfilled: R 1 + R 2 + R 3 + R 7 = R 4 + R_{tot}, where R_{tot} represents the parallel configuration of R 5 and R 6 (Both are connected to zero potential, R 6 direct and R 5 to the output where zero is usually present). However, since R 5 is substantially greater than R 6 in the case of high amplification, the following simplification is possible

$$R\,1 + R\,2 + R\,3 + R\,7 = R\,4 + R\,6 \qquad \text{Equation 3}$$

In addition to this condition, it may be necessary to provide one or two small capacitors or RC combinations in order to neutralize any tendency to RF oscillation. The values of these components are dependent on the internal build-up of the operational amplifier which means that it is impossible to give any exact values here. The data sheets and application notes of the manufacturers contain such details. Some of the available integrated operational amplifiers are built-up so that no additional circuitry will be required for most applications.

4. APPLICATIONAL NOTES

It can be seen that the accuracy of the resulting attenuation curves is solely dependent on the accuracy with which the frequency-determining components coincide to the calculated values. It is therefore important that only low tolerance resistors (5% or better) and capacitors are used.

By connecting low-pass and high-pass filters in series, it is possible to obtain a bandpass characteristic.

In order to ensure that the attenuation curve is not distorted, it is necessary to feed the filter from a source impedance that is at least ten times smaller than R 1, whereas the load impedance Z_L at the output must not be less than Z_L = 30 x R 1/B where B is the static current amplification of the transistor.

5. REFERENCES

(1) D. E. Schmitzer: NF-Filter ohne Spulen
Das DL-QTC 33 (1962), H. 3, Pages 104-107

(2) Mc Vey: An Active RC-Filter Using Cathode Followers
Electronic Engineering (1962), July, Pages 458-463

(3) D. E. Schmitzer: Preamplifiers to Improve Speech Intelligibility Under Poor Operating Conditions
VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 110-114.

ACTIVE AUDIO FILTERS
PART II - PRACTICAL CIRCUITS

by D.E. Schmitzer, DJ 4 BG

INTRODUCTION

The introductory article "Active Audio Filters" (1) is now to be extended by a series of proved and measured circuits designed to provide favourable AF response for both transmit and receive applications.

The measured frequency response curves are given for each of the given circuits. A universal printed circuit board has been developed by the author that allows all described circuits to be built up in a simple and space saving manner.

The intelligibility of processed speech is not noticeably less than wideband transmissions under good operating conditions (high signal-to-noise ratio); however, under poor conditions (high noise level, interference etc.) the advantages are very apparent. These advantages were explained in detail in (1) and (2).

1. AF FILTERS IN THE TRANSMIT MODULATOR

The available transmit energy of a voice transmitter is better utilized if the audio frequency range provided by the microphone is limited to that frequency spectrum required for voice transmission. The lower -3 dB frequency limit need not be lower than 300 Hz and the upper limit not greater than 3 kHz. In addition to this, the remaining transmission range should fall by 6 dB per octave in the lower frequency direction (half the frequency = half the gain).

1.1. SIMPLE LOW-PASS FILTER

The easiest manner of suppressing the unwanted higher frequency components above 3 kHz is to use a simple low-pass filter such as that given in Fig. 1a. The disadvantage of this circuit is that it must be driven from a low impedance source ($Z \leq 500\ \Omega$) so that the frequency response curve is not distorted. Since such a source (microphone) is not always available, the high impedance circuit given in Fig. 1b - where the source impedance may be up to 10 kΩ - could be of advantage. Smaller capacitance values also result that may be somewhat easier to obtain and which therefore tend to compensate for the extra expense of the second transistor.

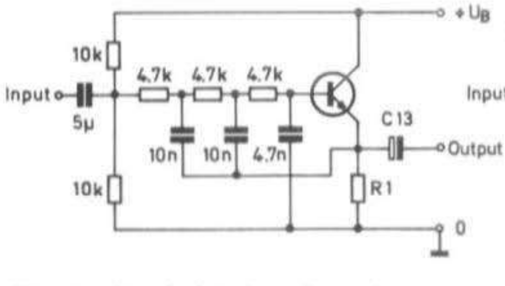

Fig. 1a Simple low impedance low-pass filter

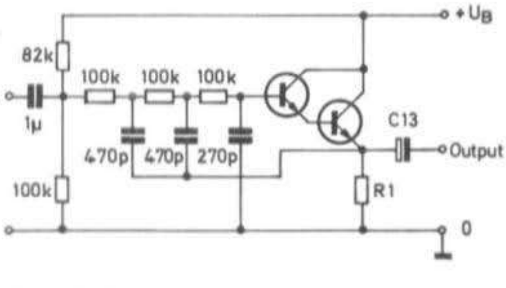

Fig. 1b Simple high impedance low-pass filter

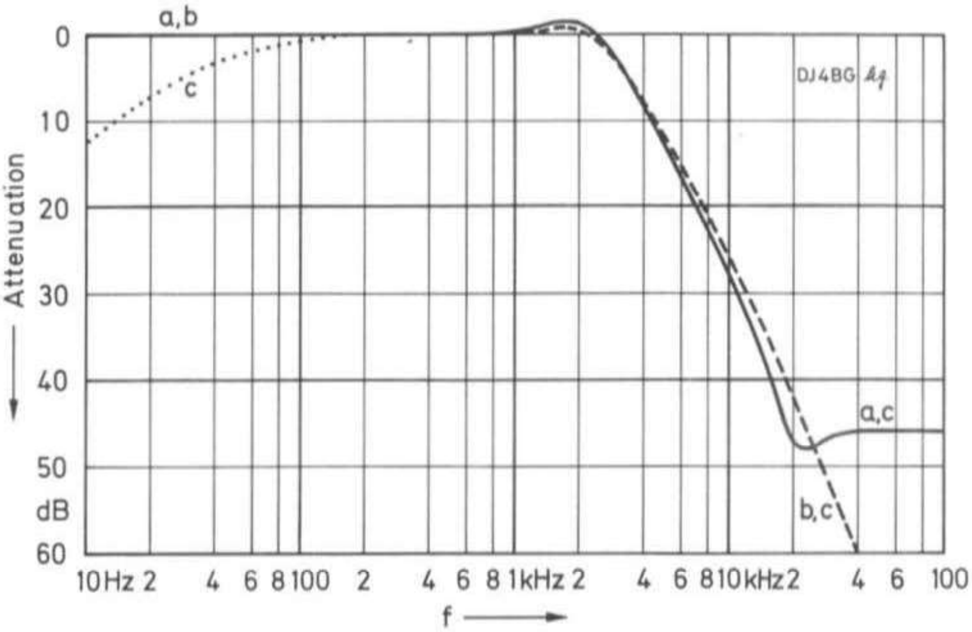

Fig. 1c Circuit 1a (curve a) and 1b (curve b) as well as circuit 2a/2b (curve c)

The measured frequency response curves of both circuits are given in Figure 1c. They were measured at an operating voltage of 9 V using the component values given in Table 2. Since the curves of circuits designed for other operating voltages offered no noticeable deviation, no further curves are given.

It can be seen that the high impedance circuit (Fig. 1b) has a more favourable response at higher frequencies. The reason for this was given in (1). In spite of this, the frequency response does not exhibit any considerable difference between the low impedance and the high impedance configuration. Since this is also valid for the following circuits, only the frequency response curve of the low impedance circuits will be given.
Both versions of the simple low-pass filter can be built up on part "B" of the printed circuit board DJ 4 BG 001 (see Fig. 7). The associated component location plans are given in Figures 1d and 1e. The values of the designated components are listed in Table 2.

1.2. ADDITION OF AN AMPLIFIER STAGE

It is possible with a suitable modification of the circuit to combine an amplifier with a low-pass filter to form a space and component saving combination. This is explained in conjunction with the simple low-pass filter in Figure 2a or 2b. The frequency response curves correspond at medium and high frequencies exactly to those of the simple low-pass filters without amplifier (Fig. 1c). At lower frequencies, the capacitors C 3 and C 4 of the amplifier stage cause a gain reduction (curve "c" in Fig. 1c). This "bass cut-off" is very desireable for speech applications. If the gain reduction may commence at 300 Hz, the capacitance values can be reduced to 1 μF for C 3 and 25 μF for C 4.

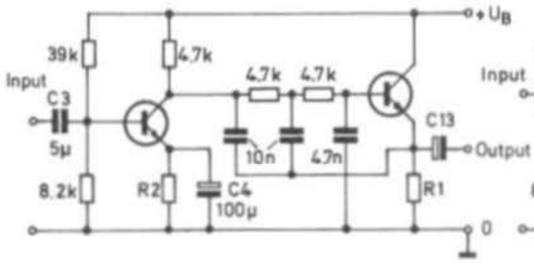

Fig. 2a Low impedance low-pass filter with amplifier

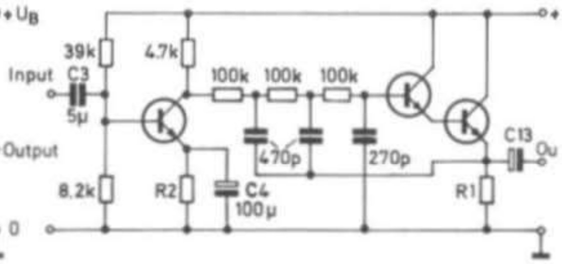

Fig. 2b High impedance low-pass filter with amplifier

The 0 dB point of the frequency response curve corresponds to the voltage amplification A in Table 1 for the circuits given Fig. 2a and 2b.

Operating voltage	6 V	9 V	12 V	18 V
Voltage amplification A for C 4 = 0 μF	7.8 17.8 dB	4.4 12.8 dB	3.9 11.8 dB	3.1 9.8 dB
Voltage amplification A for C 4 = 100 μ F	104 40.2 dB	137 42.7 dB	177 45 dB	230 47.2 dB
Current requirements	approx. 1.5 mA	approx. 2.1 mA	approx. 2.4 mA	approx. 3.2 mA

Table 1 Measured values for circuits given in Fig. 2a and 2b.

The component location plans for the low-pass filter circuits according to Fig. 2a and 2b are given in Fig. 2c and 2d.

The filters described in the following sections can also be combined with an amplifier stage providing that the input circuit is in the form of a low-pass filter. The important point is that the first filter resistor of the low impedance configuration is also used as collector resistor for the preceding transistor.

1.3. A STEEP SKIRTED LOW-PASS FILTER WITH BUFFER

If the principle is taken a little further, it is possible to build up a circuit that allows a steeper treble cut-off. The circuit, which is given in Figures 3a (low Z) and 3b (high Z), consists of a common collector stage followed by two low-pass filter sections. The corresponding frequency response curves are given in Fig. 3c and the component location plans in Fig. 3d and 3e.

If the second low-pass filter is deleted, a simple low-pass filter circuit will result with a frequency response according to Fig. 1c. The additional buffer, however, allows the low impedance circuit to be used with source impedances up to 5 kΩ.

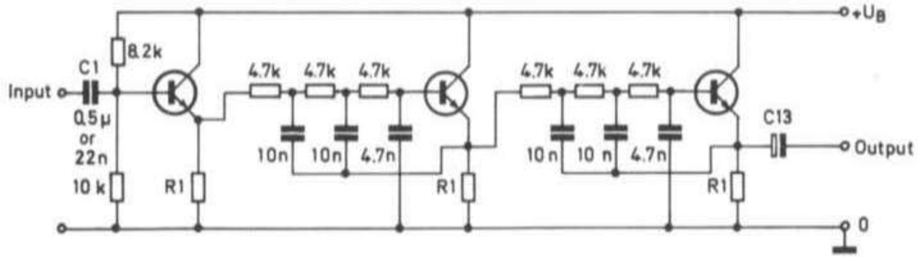

Fig. 3a Steep skirted, low impedance low-pass filter with buffer DJ 4 BG *hy*

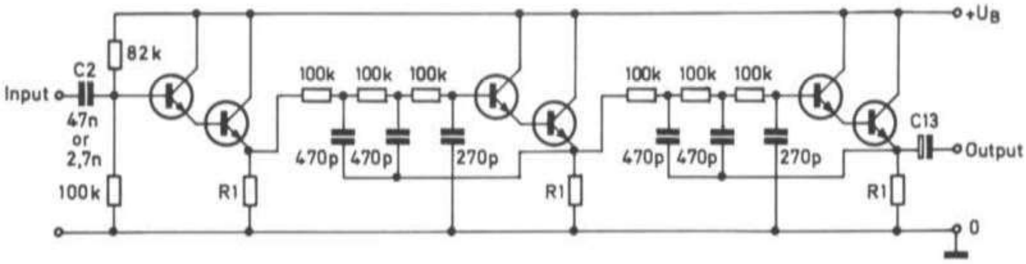

Fig. 3b Steep skirted, high impedance low-pass filter with buffer DJ 4 BG *hy*

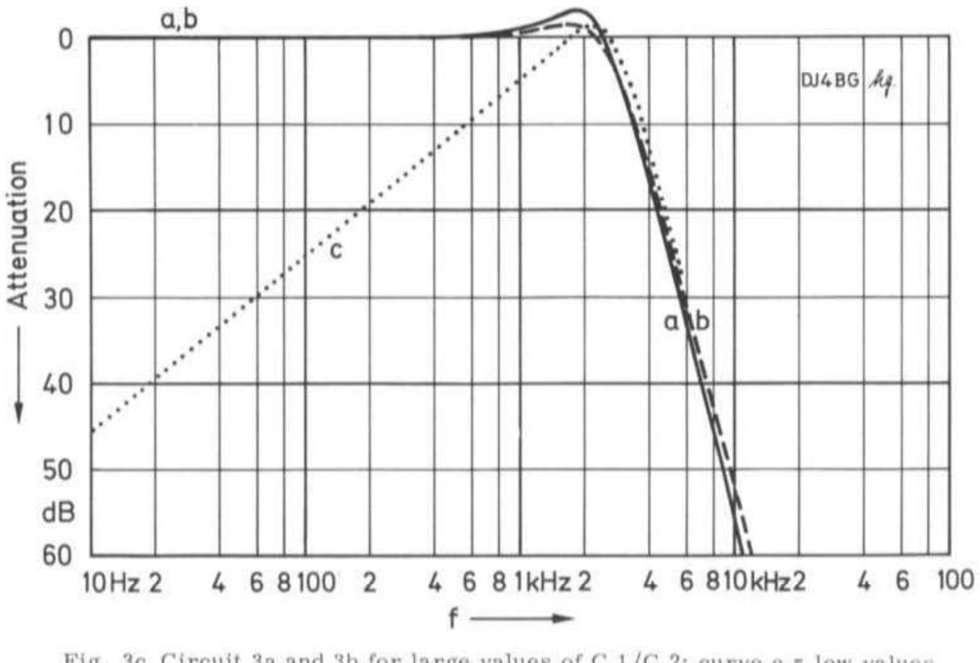

Fig. 3c Circuit 3a and 3b for large values of C 1/C 2; curve c = low values

1.4. SPEECH WEIGHTING FILTER

If the optimum frequency response for voice transmissions with a bass cut-off of 6 dB per octave is to be achieved, a point in the circuit having a defined impedance is required so that the required frequency response can be obtained using a suitable coupling capacitance. A common collector stage is suitable for this purpose since the transistor input impedance is so high that the effective input impedance is only determined by the desired low impedance of the base voltage divider. It is only necessary to use the lower values given in Table 2c for C 1 and C 2. This filter can be combined with a simple low-pass filter as shown in Fig. 4a (low Z) and Fig. 4b (high Z). Figure 4c gives the frequency response curves of the arrangement, which can be built up on part A of the PC-board. The corresponding component location plans are given in Figures 4d and 4e.

The steep skirted low-pass filter given in Fig. 3a and 3b can be used in the same manner. The frequency response curve obtained with such a configuration is also given in Fig. 3c. Both the simple and steep skirted speech weighting filter must be driven from a low impedance source; the source impedance for the low impedance configuration should be less than 1 kΩ, for the high impedance circuit less than 5 kΩ.

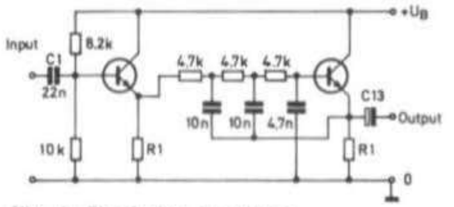

Fig. 4a Simple low impedance speech weighting filter

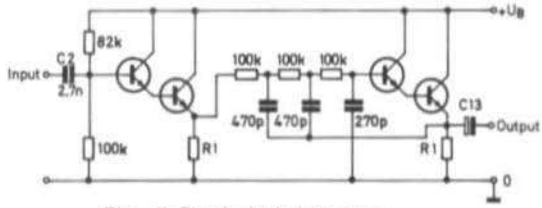

Fig. 4b Simple high impedance speech weighting filter

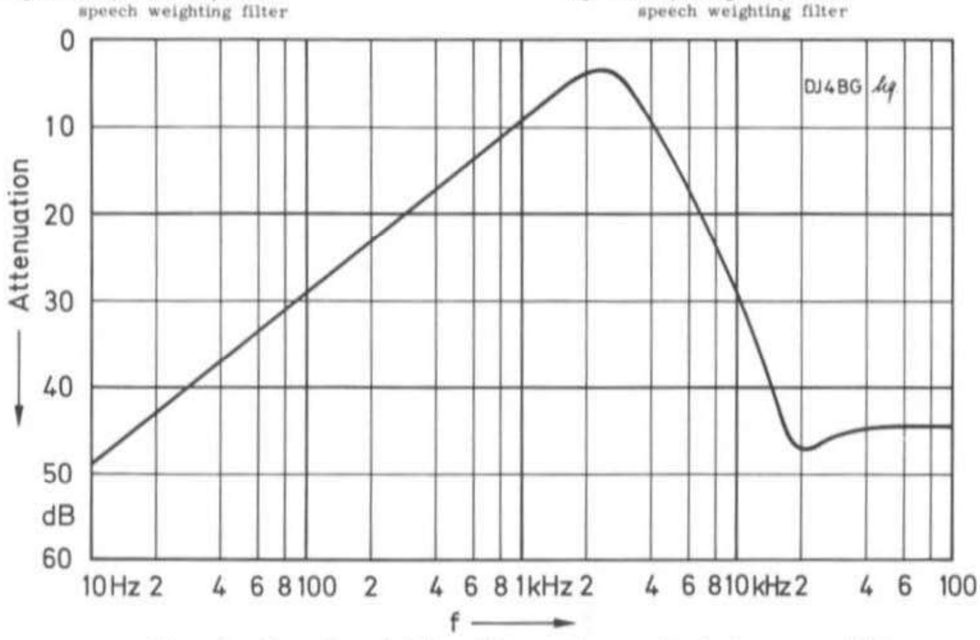

Fig. 4c Speech weighting filter using a simple low-pass filter

2. AF FILTERS FOR RECEIVER APPLICATIONS

In many cases, it is not necessary to provide additional filters in the AF chain of a receiver since the bandwidth is sufficiently narrow due to the IF selectivity. In spite of this, it is possible that additional filters could provide an improved signal-to-noise ratio. This is the case, for instance, if the IF bandwidth is essentially greater than 5 kHz, or if the overall selectivity of the receiver takes place near the input of the IF chain, perhaps with a crystal filter, and the subsequent IF stages contribute a noticeable noise component.

AF filtering is especially necessary in the receiver during reception of frequency modulated signals since only then is it possible to utilize the advantage of this mode to the full. With frequency modulation, the unwanted IF noise is no longer equally distributed after demodulation but increases its level with frequency deviation. The AF signal-to-noise ratio will therefore be improved if the frequencies above approximately 3 kHz are suppressed. This can be achieved with the filter circuits given for transmit applications in Figures 1a, 1b, 3a and 3b.

2.1. AF BANDPASS FILTERS

Since frequency components under 300 Hz do not contribute to speech intelligibility, they may also be suppressed using the bandpass filter circuits given in Figures 5a or 5b; the frequency response curve of these configurations is given in Fig. 5c. These circuits may be so dimensioned that the remaining bandwidth amounts to only about 500 Hz. This results in a considerable improvement of the signal-to-noise ratio during telegraphy reception using a receiver not equipped with a narrow band CW filter. The affected components are given in the "telegraphy filter" column of Table 2. The frequency response curve of the modified filter is given in Fig. 5c; the component location plans of the bandpass filter circuits are given in Figures 5d and 5e.

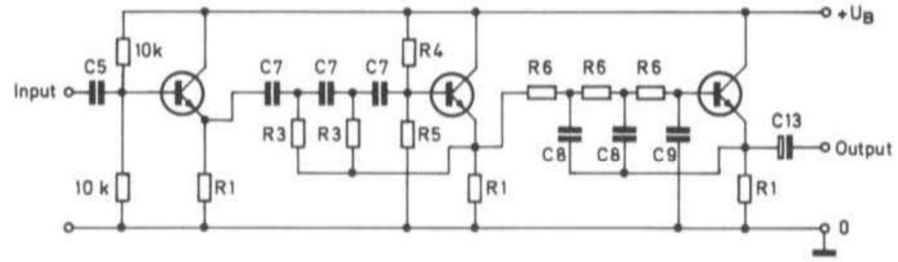

Fig. 5a Low impedance bandpass filter DJ4BG

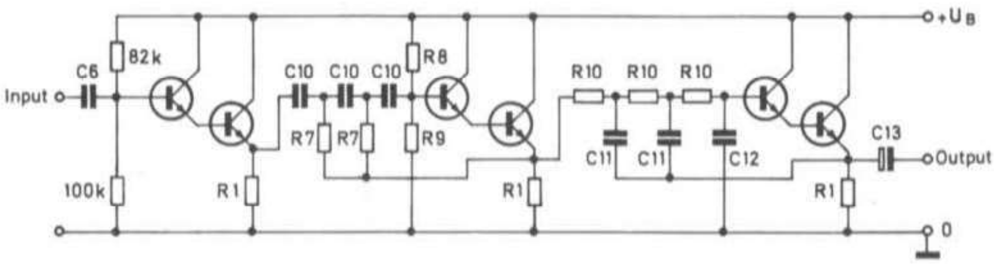

Fig. 5b High impedance bandpass filter DJ4BG

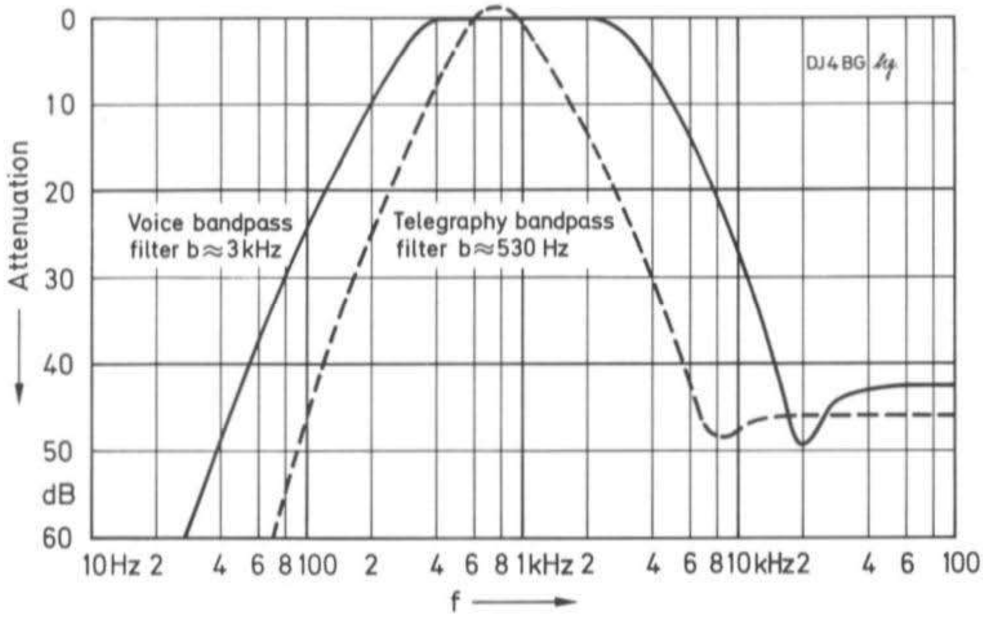

Fig. 5c Bandpass filter according to Fig. 5a or 5b

2.2. BANDWIDTH SELECTION

If the AF bandwidth of a receiver is to be switched, it will be advisable to use separate filters for telephony and telegraphy. The arrangement shown in Fig. 6 is very effective since the telephony filter increases the ultimate selectivity during reception of telegraphy.

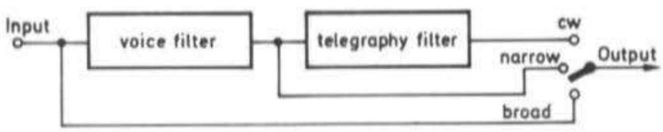

Fig. 6 Universal PC-board for active audio filters DJ 4 BG 001

3. DRIVE CAPABILITY

Generally speaking, the given filter circuits are insensitive to overload. Peak-to-peak levels up to half the operating voltage U_B can be handled without distortion which corresponds to an RMS input voltage of 1/6th of U_B. When using a preamplifier, the permissible input voltage will be reduced by the value of the gain factor.

4. COMPONENTS

The author used the silicon transistor BC 167 for all circuits. The types BC 168, BC 169 and the parallel typec BC 107 to BC 109 as well as any silicon NPN transistors having a high current amplification ($B \geq 100$ at $I_C = 1$ mA) are equally suitable. Especially low-noise transistors will only be required in conjunction with extremely low levels.

All resistors and capacitors should have the lowest tolerance possible (5% or less). Although good frequency response curves can be obtained with components having higher tolerance values, certain deviations from those given in this description will have to be expected.

The following table (Table 2) list the values of the components given in the circuit diagrams:

U_B =	6 V	9 V	12 V	18 V
R 1 =	2.7 kΩ	3.9 kΩ	5.6 kΩ	8.2 kΩ
R 2 =	560 Ω	1 kΩ	1.2 kΩ	1.5 kΩ

Table 2a Component values dependent on the operating voltage U_B

	without bass rejection	with 6 dB/octave
C 1	0.5 μF	22 nF
C 2	47 nF	2.7 nF

Table 2c Input capacitors

	Voice	Telegraphy
R 3	56 kΩ	33 kΩ
R 4	220 kΩ	120 kΩ
R 5	270 kΩ	180 kΩ
R 6	4.7 kΩ	6.8 kΩ
R 7	560 kΩ	330 kΩ
R 8	2.2 MΩ	1.2 MΩ
R 9	2.7 MΩ	1.8 MΩ
R 10	47 kΩ	68 kΩ
C 5	1 μF	0.1 μF
C 6	0.1 μF	10 nF
C 7	10 nF	10 nF
C 8	10 nF	22 nF
C 9	5 nF	10 nF
C 10	1 nF	1 nF
C 11	1 nF	2.2 nF
C 12	500 pF	1 nF

C 13 according to the impedance of the following stage: between 0.1 μF and 10 μF

Table 2b Values of the bandwidth determing components

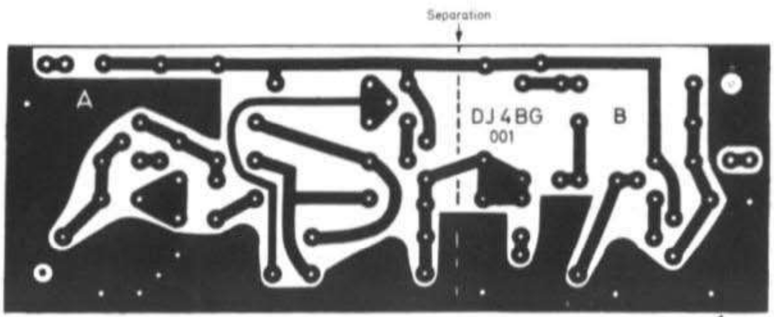

Fig. 7 Universal board for active audio filters

4.1. PRINTED CIRCUIT BOARD DJ 4 BG 001

In order to facilitate assembly, a printed circuit board having the dimensions 100 mm x 35 mm was designed on to which all the described filter circuits can be built. Due to the versatility of the PC-board, it will be necessary to leave some component positions vacant or to make individual wire bridges to suit each of the circuits. The PC-board can be separated at the dotted line (see Fig. 7) for simple configurations so that further variations are possible. The conductor side of this printed circuit board is given in Figure 7; the corresponding component location plans are given in Fig. ...d and ...e of each individual circuit.

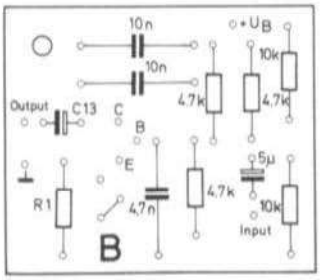

Fig. 1d Component location plan to Fig. 1a

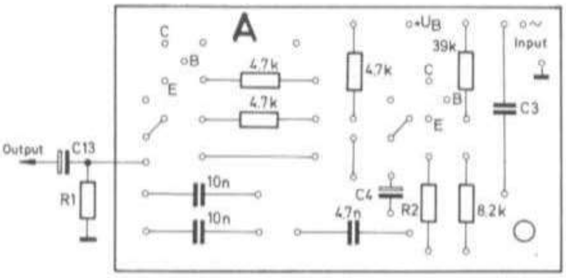

Fig. 2c Component location plan to Fig. 2a

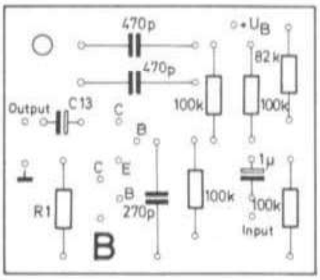

Fig. 1e Component location plan to Fig. 1b

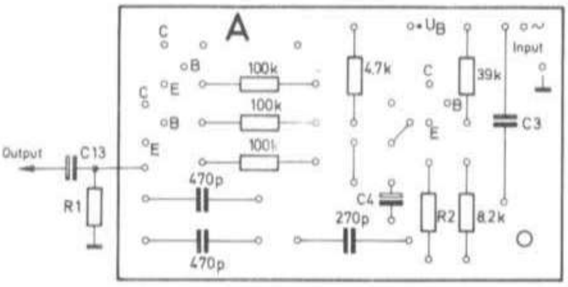

Fig. 2d Component location plan to Fig. 2b

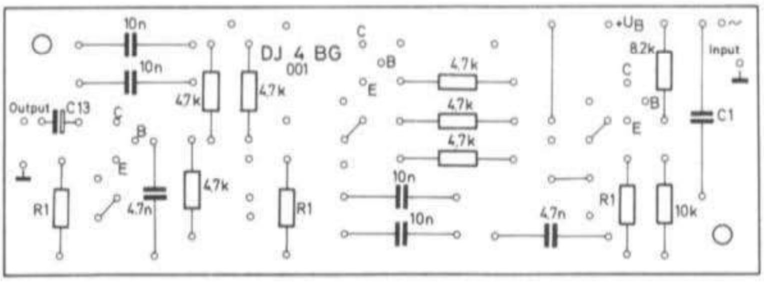

Fig. 3d Component location plan to Fig. 3a

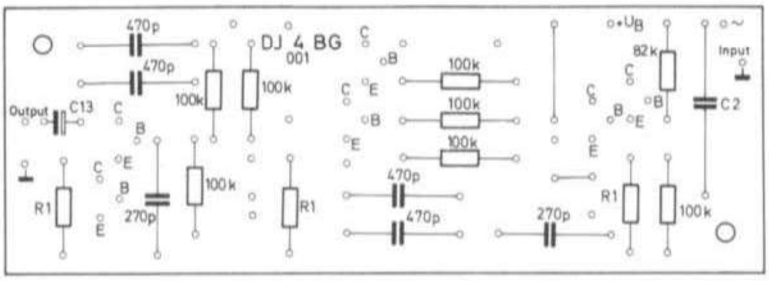

 Fig. 3e Component location plan to Fig. 3b

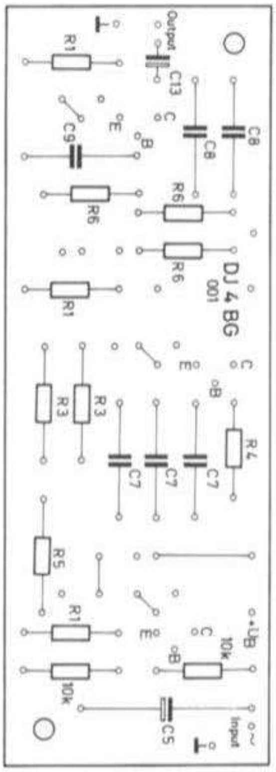

Fig. 5d Component location plan to Fig. 5a

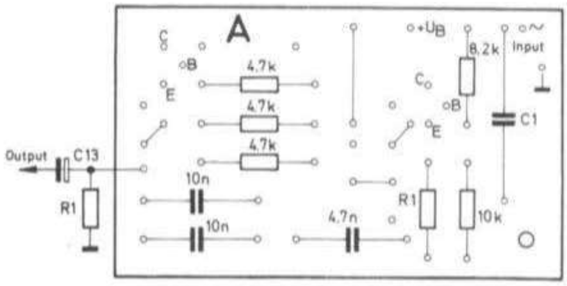

Fig. 4d Component location plan to Fig. 4a

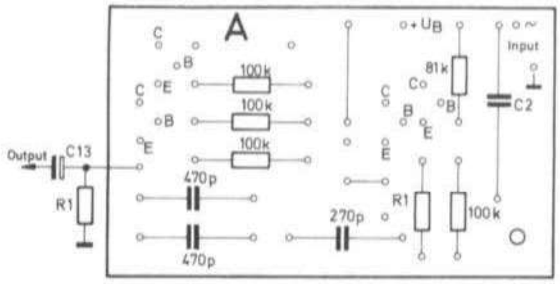

Fig. 4e Component location plan to Fig. 4b

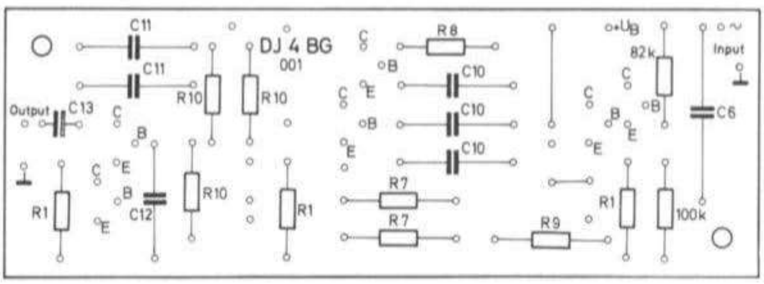

Fig. 5e Component location plan to Fig. 5b

4.2. AVAILABLE COMPONENTS

The printed circuit board DJ 4 BG 001 is available from the publishers or their national representatives. See advertising page.

5. REFERENCES

(1) D. E. Schmitzer: Active Audio Filters Part I
VHF COMMUNICATIONS 1 (1969), Edition 4

(2) D. E. Schmitzer: Preamplifiers to Improve Speech Intelligibility under Poor Operating Conditions
VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 110 to 114.