potential ECR also decreases, thus increasing $V_{in} - V_{out}$ and further decreasing the value of C in a cumulative manner. Although a critical voltage V_1 can no longer be defined accurately, it can be assumed that the change, once initiated, will be rapid.

Fig. 2b shows the output from a nonlinear differentiator employing a BA110 diode. In this circuit, the input and output voltages are related by

$$V_{out} = RC_{diode}\frac{d(V_{out} - V_{in})}{dt}$$

A computer, programmed to solve this equation by the Runge–Kutta method, produced the output waveform shown in Fig. 1f, which is in reasonable agreement with the experimental curve. Such a circuit has been used successfully to produce short-duration pulses from avalanche transistors.[2]

Another aspect of this circuit is its ability to minimise the effect of varying input amplitudes, as illustrated in Fig. 2c.

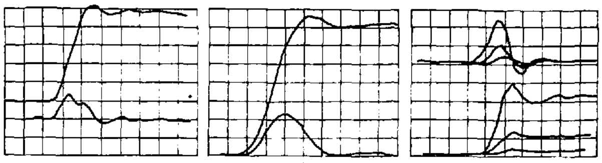

Fig. 2

a Nonlinear differentiator using a 2N2378 f.e.t. and 50Ω resistor
Top trace: input, 2V/division; Lower trace: output, 0·2V/division } 2ns/division

b Nonlinear differentiator using a BA110 diode and a 90Ω resistor
Top trace: input, 2V/division; Lower trace: output, 2V/division } 1ns/division

c Effect of amplitude variations in a nonlinear differentiator
Top trace: outputs, 0·4V/division; Lower traces: inputs, 4V/division } 10ns/cm

The range of input-pulse amplitudes is 20 : 1, but the output range in amplitudes is only 5 : 1. For small-amplitude inputs, the diode behaves as a large capacitance, but for large amplitude inputs the diode capacitance quickly decreases to a lower value, and the output amplitude is restricted. This standardisation of output-pulse amplitude has been used to advantage to obtain improved triggering pulses in high-speed scalers.[3]

This circuit has also been used to produce very short-duration pulses from nonlinear delay lines.[4]

Nonlinear differentiators have been analysed by other workers for step-function inputs,[5] and similar circuits containing inductances have also been analysed.[6]

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DIRECT-COUPLED GYRATOR SUITABLE FOR INTEGRATED CIRCUITS AND TIME VARIATION*

A direct-coupled gyrator circuit which can be easily realised in integrated-circuit form is described. Since the gyrator action extends to d.c., it is practicable to realise d.c. transformers by cascading two such gyrators.

Practical gyrator circuits that do not depend upon cancellation of transistor parameters by negative resistors, in contrast to other gyrator circuits,[1–3] have been presented in References 4, 5 and 6. These circuits require blocking capacitors at the input and output, which are not practicable at low frequencies. Consequently, a direct-coupled gyrator that can be easily integrated has been developed. Since the gyrator action extends to d.c., it is practicable to realise d.c. transformers by cascading two such gyrators.

The operation of the circuit is best understood from the block diagram in Fig. 1, which consists of two differential voltage/current convertors with high input and output impedances connected back to back, as shown. By definition, each differential convertor has two voltage inputs, with the difference of the two voltages determining the output current.

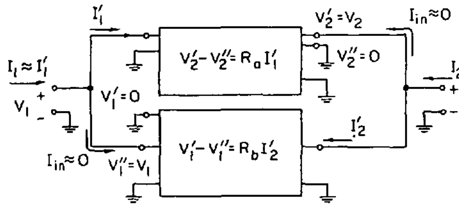

Fig. 1 *Gyrator as differential voltage/current convertors*

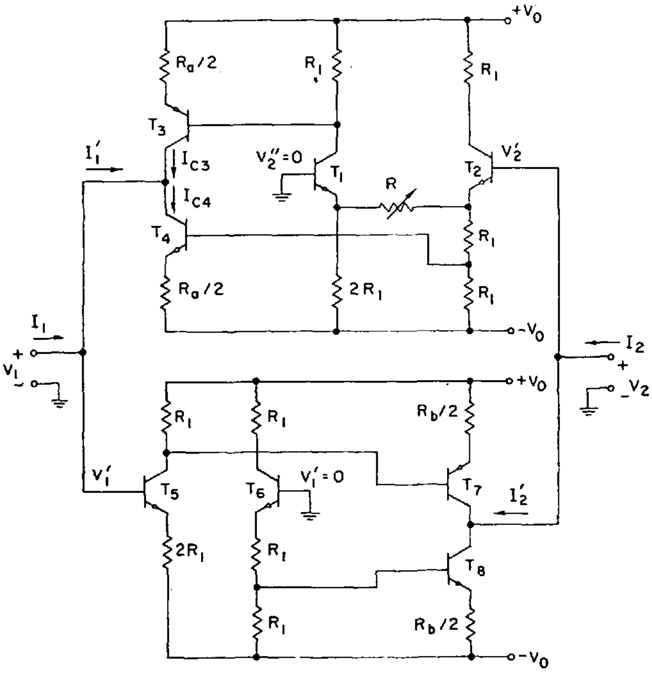

Fig. 2 *Transistor realisations*

On grounding complementary inputs in the convertors, the connection of Fig. 1 immediately gives

$$V_1 = -R_bI_2 \quad (1a)$$

$$V_2 = R_aI_1 \quad (1b)$$

and hence, when $R_a = R_b$, the circuit is a gyrator with gyration resistance R_b; even when $R_a \neq R_b$, it is convenient to call the device a 'gyrator'.

In the actual circuit shown in Fig. 2, transistors T_1–T_4 and the associated resistors form the upper differential voltage/current convertor, while T_5–T_8 form the lower convertor. The resistor R is for time variation and, for preliminary understanding, can be assumed infinite at this

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point. The base of T_1 is at ground potential ($V_2'' = 0$) at all times. With zero voltage applied to the base of T_2, it can be seen that the indicated collector currents I_{C3} and I_{C4} are exactly equal, and $I_1' = (I_{C4} - I_{C3})$ is zero. When $V_2' = V_2$ is nonzero, the variation in I_{C4} from its quiescent value is given by V_2/R_a, since V_2 is halved by the two resistors at the emitter of T_2 and therefore appears across the emitter resistor of T_4, causing I_{C4} to vary, whereas I_{C3} remains at a constant value equal to the quiescent value of I_{C4}. This results in $I_1' \simeq I_1$, which is eqn. 1*b*. The operation of the lower circuit is, of course, identical.

At low and medium frequencies, the input and output impedances of each of the differential voltage/current convertors are essentially resistive, and the parallel combination of these may be taken into account by considering parasitic resistors shunting each port of an ideal gyrator. When this nonideal gyrator is terminated at port 2 by a capacitor C, it can be shown that the input impedance seen at port 1 looks like an inductor of essentially constant value of R_b^2C, up to an angular frequency of about R_p/R_a^2C, where R_p is the parasitic resistor. This equivalent inductance has a maximum Qfactor of $R_p/2R_a$ at an angular frequency of $1/R_aC$. The Qfactor of the effective inductor increases from zero to a maximum value and then goes down, making the input behave more like a regular coiled-wire inductor than an ideal inductor.

The gyration resistances in eqn. 1 can be varied in several ways. One is to vary the correspondingly labelled resistors in Fig. 2, say, by varying the source–drain resistances of a field-effect transistor.[4] Since this requires simultaneous variation of two resistors for varying either R_a or R_b, one also notes that R_a of eqn. 1*b* (but not R_a of Fig. 2) can be varied as follows. If one inserts the resistor R between the emitters of T_1 and T_2, R_a in eqn. 1 is replaced by $R_a/\{1 + (4R_1/R)\}$, which then varies with the single resistor R. In this, R_b in eqn. 1 remains constant, but it can similarly be varied if so desired.

To increase the input and output impedances of the differential voltage/current convertors, each of the transistors T_1–T_2, T_7–T_8 may be replaced by Darlington pairs. Also, the simple differential voltage/current convertors used here can be replaced by more sophisticated ones with internal and/or external feedbacks to increase input and output impedances.

Finally, we acknowledge the assistance of P. Salsbury and P. Gary, who have constructed and tested a time-invariant lumped model appropriate for integration (with $V_0 = 9$V, $R_1 = 680\Omega$, $R_a = R_b = 2400\Omega$, $R_p = 300\,\text{k}\Omega$). Experimental verification agrees excellently with predicted behaviour and will soon be available.[7] Similar time-invariant devices have also been designed by Hawley[8] and Sheahan;[9] these also exhibit excellent performance.

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WAVEGUIDE BANDPASS FILTERS USING EVANESCENT MODES

Bandpass filters using transmission lines with a positive imaginary characteristic impedance, terminated in a capacitive reactance, are described. Physical realisation takes the form of waveguide beyond cutoff (H_{01} mode) and, typically, a capacitive tuning screw. The screw is the only obstacle required in the guide.

The commonplace situation in which a transmission line of real characteristic impedance is terminated in a load, equal to its impedance, is the basis of much of microwave technology. It is, perhaps, not so obvious that interesting possibilities occur when a transmission line of pure imaginary characteristic impedance is suitably terminated. A transmission line of the latter type will have a real propagation coefficient, which, for significant lengths, can introduce appreciable attenuation (by reflection) between source and load. However, if the line is terminated in its conjugate impedance, full energy transfer takes place. The energy transfer is usually highly frequency-sensitive, and the device behaves as a bandpass filter. Such a filter is represented by the circuits of Fig. 1.

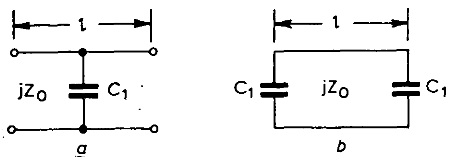

Fig. 1 *Equivalent circuit of filter*

Analysis on an image-parameter basis shows that over a range of frequencies the image impedance of the filter is real (the passband). The limit frequencies f_1 and f_2 at which the image impedance is, respectively, infinity and zero are given by

$$f_1 = \frac{\tanh \frac{1}{2}\gamma l}{2\pi Z_0 C_1} \qquad (1)$$

$$f_2 = \frac{\coth \frac{1}{2}\gamma l}{2\pi Z_0 C_1} \qquad (2)$$

The centre frequency $f_0 = \sqrt{(f_1 f_2)}$ is given by

$$f_0 = 1/2\pi Z_0 C_1 \qquad (3)$$

The image impedance Z_{I0} at this frequency is given by

$$Z_{I0} = Z_0\sqrt{(\tanh \tfrac{1}{2}\gamma l)} \qquad (4)$$

This image impedance is, of course, real and, for reflectionless operation at the centre frequency, must be matched by further sections of identical impedance or a propagating line of the same real characteristic impedance.

Physical realisation of the filter involves waveguide beyond cutoff, which for Hmodes is a pure inductive reactance. A single Tsection filter (Fig. 2) then requires a capacitive

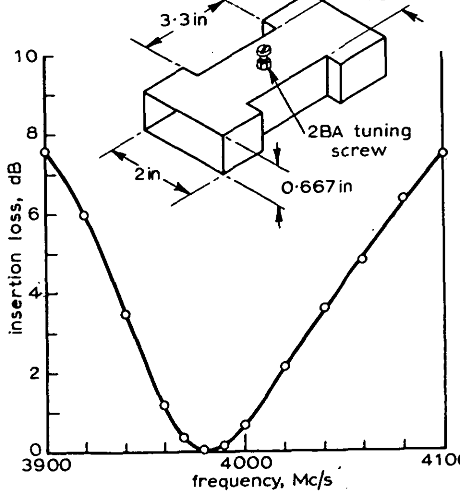

Fig. 2 *Performance of one Tsection filter*